AD-A058 367

AD-A058 367

DNA 2772T

ADA 058-367

DNA EMP AWARENESS COURSE NOTES

Third Edition

IIT Research Institute
10 West 35th Street
Chicago, Illinois 60616

October 1977

Topical Report

CONTRACT No. DNA 001-75-C-0074

APPROVED FOR PUBLIC RELEASE;
DISTRIBUTION UNLIMITED.

THIS WORK SPONSORED BY THE DEFENSE NUCLEAR AGENCY
UNDER RDT&E RMSS CODE X323076469 Q76QAXEC09201 H2590D.

Prepared for
Director
DEFENSE NUCLEAR AGENCY
Washington, D. C. 20305

35

UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER DNA 2772T	2. GOVT ACCESSION NO.	3. RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) DNA EMP AWARENESS COURSE NOTES Third Edition,		5. TYPE OF REPORT & PERIOD COVERED Topical Report 6. PERFORMING ORG. REPORT NUMBER
7. AUTHOR(s) I. N. Mindel		8. CONTRACT OR GRANT NUMBER(s) DNA 001-75-C-0074
9. PERFORMING ORGANIZATION NAME AND ADDRESS IIT Research Institute 10 West 35th Street Chicago, Illinois 60616		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS NWED Subtask Q75QAXEC092-01
11. CONTROLLING OFFICE NAME AND ADDRESS Director Defense Nuclear Agency Washington, D.C. 20305		12. REPORT DATE October 1977 13. NUMBER OF PAGES 278
14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)		15. SECURITY CLASS (of this report) UNCLASSIFIED 15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited.

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

18. SUPPLEMENTARY NOTES

This work sponsored by the Defense Nuclear Agency under RDT&E RMSS Code X323075469 Q75QAXEC09201 H2590D.

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Electromagnetic Pulse	EMP Design Practices
EMP Environment	EMP Testing
EMP System Degradation Modes	EMP System Hardening
EMP Interaction and Coupling	

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

These notes are primarily intended as a guide for the individual attending the DNA EMP Awareness Course. The text serves as an engineering introduction to the EMP systems desi problem and provides a survey of techniques suitable for solv. this hardening problem. The text discusses all aspects of EMP, from the environment through system design and testing.

SUMMARY

This course is designed to alert management and technical personnel to the need for EMP protection and to the hardening techniques which can be applied. The bulk of the course material is aimed toward those individuals who may have a direct responsibility for EMP hardening of a system.

The duration of the complete course is three days. The course also includes an Overview of interest to management.

The Overview is essentially an introduction to the EMP problem area. The general problems and difficulties are reviewed.

The technical portion of the course considers all aspects of the problem area from an engineering hardening viewpoint. It requires approximately 2-3/4 days of presentation time to discuss the salient features of the information and background necessary to harden systems.

PREFACE

Third Edition

To maintain the course and the course notes current, updating and revisions are a continuous effort. Since the original publication in August 1971, significant advances in the state-of-the-technology have occurred. In that regard, complete revision of the course notes was deemed desirable. The technical revision was performed by Mr. Irving N. Mindel of IIT Research Institute.

The author of the course notes would like to acknowledge the leadership and technical review of the course and the course notes provided by Captain W. D. Wilson and Mr. G. Baker from the Defense Nuclear Agency. Acknowledgement of the comments received from the "EMP Community" and the course attendees has also been very beneficial to furthering the objectives of the course. The efforts of the editorial staff of IITRI, particularly Ms. M. Leek and Ms. C. Damberger, who compiled and edited the final publication, is gratefully acknowledged.

I. N. Mindel
IIT Research Institute
Chicago, Illinois
October 1977

TABLE OF CONTENTS

TABLE OF CONTENTS

TABLE OF CONTENTS

TABLE OF CONTENTS

TABLE OF CONTENTS

SECTION I

OVERVIEW:

INTRODUCTION TO THE EMP PROBLEM

1.1 INTRODUCTION

Everyone is aware of three major effects of a nuclear weapon; these are blast, thermal, and shock. There are other effects associated with a nuclear detonation, which are equally important. One of these effects is the electromagnetic pulse (EMP) which is the subject of this awareness course.

Planners now consider EMP as a possible threat to many sophisticated military systems. How this has come about and, especially, how to achieve the required protection will be the subject of this introductory portion and the more detailed technical presentations which follow. These technical presentations emphasize the engineering aspects.

Under the proper circumstances a significant portion of the energy released during a nuclear detonation appears as an ElectroMagnetic Pulse (EMP). The spectral content of this pulse contains the frequencies or wavelengths of many of our commercial radio and military electronic systems.

Another nuclear weapon effect sometimes confused with EMP is "blackout". As we will see later, EMP directly affects electrical and electronic systems. "Blackout" on the other hand, is due to an effect associated with the propagation media (the atmosphere) and disrupts the signal path. "Blackout" has no direct effect on system hardware. Also, EMP is a short-term transient (microseconds) that can result in long-term disruptive effects in equipments/systems.

EMP effects depend on several factors among which are the weapon, the burst location, and the point of observation with respect to the burst location. In this course we will be concerned with the surface, near surface and the exoatmospheric (or high altitude) burst. Emphasis will be placed on the exoatmospheric burst because it has two unique properties which are of crucial significance.

These unique properties are its extreme area of coverage, EMP being capable of disrupting electrical and electronic systems as far as 3,000 miles from the site of the detonation; and the fact that EMP can cause severe disruption and sometimes damage when other prompt weapon effects, such as nuclear radiation effects on electronics, blast, thermal effects, dust debris and biological effects are all absent. This means that a high-yield nuclear weapon, burst above the atmosphere, could be used to knock out improperly designed electrical and electronic systems over a large area of the earth's surface without doing any other significant damage. The range of coverage of EMP is diminished if the weapon is detonated within the atmosphere.

An idea of the magnitude of the EMP threat can be realized by comparing its electric field amplitude with the electric fields produced by man-made sources. A typical high-level EMP pulse could have an intensity of 500,000 volts per meter. This is 250 times more intense than a radar beam of sufficient power to cause biological damage such as blindness or sterilization. It is five million times as intense as fields created by sources in a typical metropolitan area.

ELECTROMAGNETIC FIELDS:

SOURCE	INTENSITY ([illegible])
EMP	50,000
Nearby RADAR	[illegible]
Nearby COMMUNICATIONS	10
Typical METROPOLITAN AREA	[illegible]

Early 1960's

Interest---Peripheral

Early 1970's

Central Issue: Protection

EMP has been recognized as a potential threat to our electronic and electrical systems since the 1960's. Recently two factors have greatly increased the significance of this threat:

(1) Increased sophistication in nuclear strategy and weapons.

(2) Increased susceptibility of electronic systems due to the broad introduction of semiconductors and newer electronic technologies and the ever-greater dependence on complex operational hardware.

The situation has changed so much that the central issue today is PROTECTION --

- how to protect your system against the potentially-massive disruptive logic upset type effects of EMP -- such as locking out the launch control or missile guidance subsystems during a critical period, etc; --
- how to protect your system against the sometimes capricious-type EMP permanent damage -- such as destroying the front ends of HF communications receivers.

To understand how to protect our systems requires some understanding of the EMP generation mechanisms and characteristics, the susceptibility characteristics of electrical and electronic systems, and the various means to counter the EMP effects.

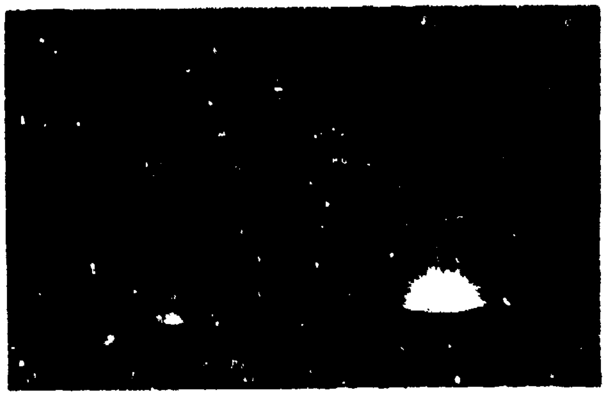
1.2 EMP GENERATION AND CHARACTERISTICS

Two basic weapon detonation locations are of interest; the near surface burst and the exoatmospheric burst. While the basic mechanism is the same, the atmospheric density results in significantly different interaction and source region characteristics.

EMP Generation

The basic mechanism for EMP generation is electron scattering due to the collision of the prompt gammas with the molecules of the media exposed to the gamma flux. When the collision is with air molecules of the atmosphere resulting in a radiated wave, we have the classic EMP. Collisions of the gamma flux or x-rays with a system enclosure (such as a satellite system), results in an EMP which is termed close-in or source region EMP. In this course we will constrain ourselves to the radiated EMP from a high altitude or surface burst.

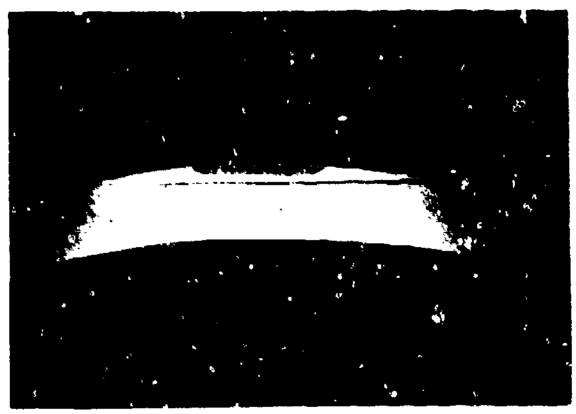
The collision between the gamma flux and the air molecules results in electrons being scattered in approximately the same direction as the gammas. The result is a charge separation, positively charged center due to the parent molecules and a negatively charged electron cloud. This separation of charges, occurring on a wholesale basis, creates intense source region electric and magnetic fields. This build-up of fields is ultimately limited by secondary conduction electrons which flow back to the positively charged region and tend to neutralize it. These source region fields are non-radiating fields.



On the other hand, if the weapon is detonated outside the atmosphere, usually termed high-altitude burst, the gamma rays can travel many miles without encountering an air molecule. Thus, if a large-yield weapon is detonated just above the ionosphere, the source region can be about 1,600 km in diameter and about 20 km thick. The extent of the electromagnetic fields radiated onto the earth's surface is greatly augmented by the very large size of this source region.

To produce a radiated electromagnetic field the proper conditions of asymmetry or interaction with the earth's geomagnetic field must exist. These conditions exist for both the high altitude and surface burst, but the dominant phenomena is different for different burst locations.

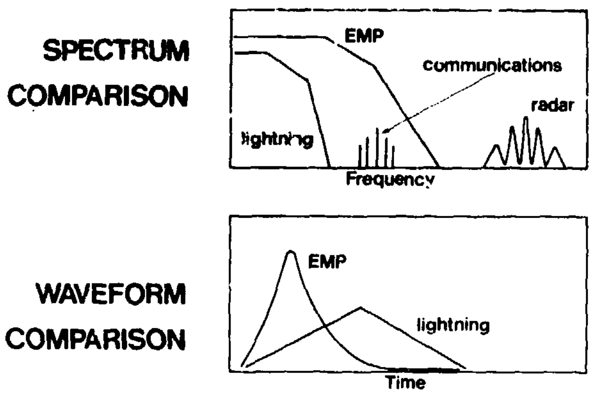
When a weapon is detonated within the atmosphere, the gammas can travel only relatively short distances between collisions and total absorption. This confines the gammas (and therefore the source region) to a volume perhaps 6 kilometers or so in diameter. Increase in weapon yield has relatively little effect upon the size of this intense field (or "source region"), and thus the EMP effects of an intra-atmospheric burst must be considered within the context of other close-by prompt nuclear weapons effects.



Characteristics of the Radiated Wave

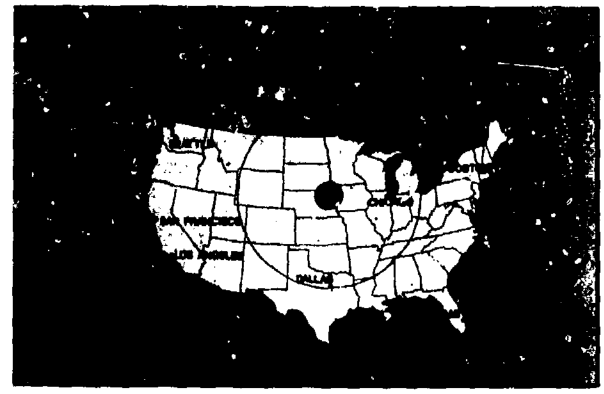
The spectrum and waveform of EMP differ from those of any other natural or commonly-used man-made sources. The spectrum is broad and extends from extremely low frequencies to the low end of the UHF band. The time waveform indicates a higher amplitude and much faster rise time than, for example, the fields generated by a nearby lightning stroke, especially in the case of high altitude burst.

A comparison of the waveform and spectrum of the high altitude burst and the induction fields of a lightning stroke are shown in the figure.



Although a lightning stroke can have a fast rise time, low energy precursor before the main stroke, the main stroke is a high amplitude (100 kV/m or greater), 1 to 5 microsecond rise time and hundreds of microseconds fall time. An enormous amount of energy is contained in the main stroke. The induction field, however, which is of concern to systems nearby a lightning discharge is on the order of 1 kV/m electric field. This induction field is non-radiating and therefore a localized field.

In the high altitude burst case, the fields radiated onto the earth's surface are of the order of 50 kV/m electric field with rise times of the order of 10 nanoseconds. The wave is a radiating EM wave which results in an extremely large distribution on the earth's surface in contrast to the localized nature of lightning. For a burst over the continental U.S., the fields on the surface are predominantly horizontally polarized E fields. Since the burst location is outside the ionosphere, the coverage on the earth's surface is limited by the line-of-sight tangent radius to the earth's surface. The figure shows the approximate coverage on the earth for a 100 km (small circle) and 300 km (large circle) height of burst over the central U.S.



These fields are not uniform in amplitude or waveform over the entire area but depend on burst location and the earth's geomagnetic field, as we will see in Section III of the course. It should also be recognized that no other weapons effects accompany the EMP from a high-altutude burst.

Due to the fast rise times of the EMP, the spectral energy is distributed throughout the spectrum through the lower microwave band. Most man-made sources occupy only a narrow part of the spectrum.

In the case of the near surface (0-2 km) burst within the deposition (source) region, non-radiating fields on the order of 100 kV/m electric fields (or higher) with rise times of tens of nanoseconds (approximately 50 ns) are realized. Outside the deposition region, a radiating EM wave is realized. The fields of this radiating wave at a distance of 10 km from the burst location are quite comparable to the close-by fields of the lightning stroke. That is, rise times of approximately 1 to 5 microseconds and peak amplitude of 1 kV/m electric field. These fields fall off as 1/R (R being the distance from the burst point) with distance from the burst. These fields are essentially vertical polarized electric fields.

In other words, EMP is sufficiently different from any other electromagnetic environment usually encountered that protection practices and components for non-EMP environments -- radio-frequency interference, lightning, radar, etc., are not directly applicable for EMP problems.

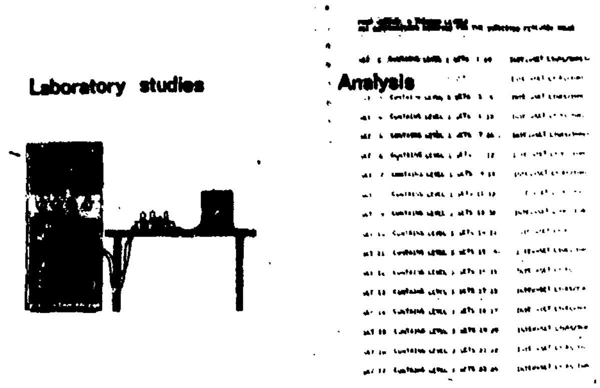
1.3 SYSTEMS IMPACT OF EMP

During the era of U.S. atmospheric testing, the existence of EMP was known, but the impact on electrical systems was not fully appreciated until some evidence of actual damage to components or system upset were accumulated. This absence of early evidence was due to relatively unsophisticated test exposure hardware (such as buildings, tanks, and jeeps) together with the subtle nature of the effects.

The most sophisticated test hardware was usually associated with diagnostic instrumentation and this did exhibit noticeable effects.

Numerous examples of EMP-induced damage or disruptions were observed during our atmospheric test programs. This photograph of a damaged cable shows an arc-type puncture. Other types of degradation attributed to EMP ranged from the tripping of a street light system to disruption of timing circuits for a communications system.

Since the signing of the atmospheric test ban treaty, the United States effort has refined the understanding of how EMP is generated to a point where good agreement exists between the theoretical results and observed measurements taken during the atmospheric tests. How EMP is coupled into systems and how components and subsystems are degraded has also received emphasis.



Effort was also devoted toward simulating experimentally the EMP field at threat levels. Even over relatively small areas, such as that occupied by a missile, this non-nuclear EMP simulation has proved to be a very difficult and costly problem because of the large amplitude and unique waveforms required. Shown, together with a missile being tested within the wooden structure, is an early (1968-1970) criteria level EMP simulator. Later generation EMP simulators and illuminating structures are discussed in Section VII.

Modern sophisticated transistorized systems have been experimentally exposed to simulated EMP environments. These tests show evidence of significant damage and extensive performance interference.

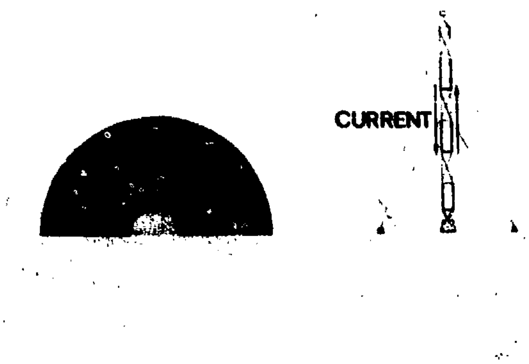
A Conclusion -- EMP is a Problem

The results developed from laboratory EMP sensitivity tests on components and electromagnetic coupling analysis further confirm this experimental data.

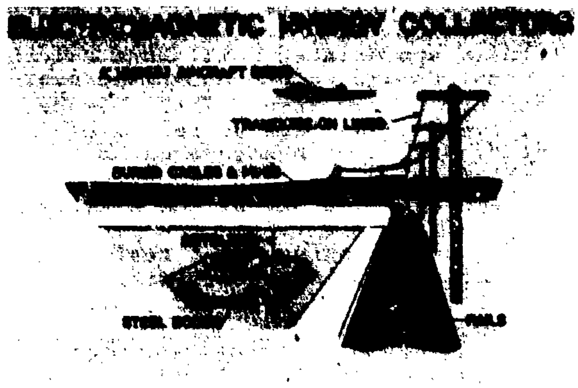
EMP Coupling to Systems

For energy to enter a system and result in performance degradation, the energy must first be collected by the system. This energy collection process can be considered to be the same as for any antenna system.

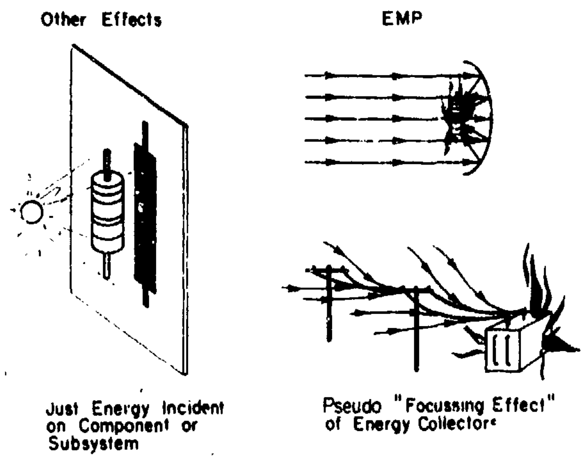
The asymmetrical flow of charges caused by a nuclear explosion causes electromagnetic fields to be radiated away from the burst point. These fields cause a corresponding flow of charges, or electrical currents, in distant metallic conductors. This is comparable to the way the electromagnetic fields from a TV transmitter set up currents in your rooftop TV antenna.



Any metallic object exposed to electromagnetic fields can be a collector of electromagnetic energy; that is, act like an antenna, even though it was never intended to be that. Generally, the larger the metallic structure, the greater is the amount of intercepted EMP energy.



Herein lies an important difference between other nuclear weapons effects (such as thermal, blast, and transient radiation) and EMP. These non-EMP effects result from just the energy incident on the sensitive component, circuit or subsystem. In the case of EMP, the energy is gathered by antennas, long exposed cables and transmission lines, or other long exposed conductors. A large fraction of this gathered energy can be impressed on a sensitive component, such as a transistor, and cause either permanent degradation (burnout) or false responses (upset) of a system.

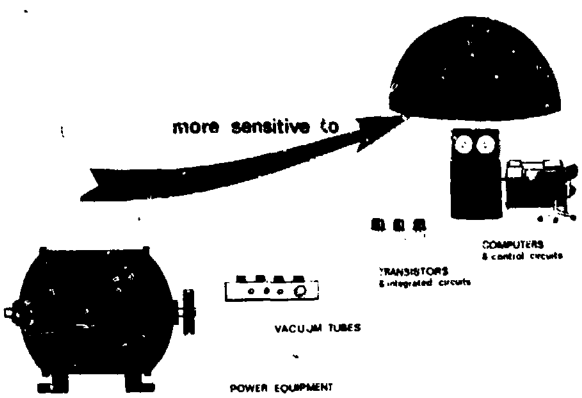


System Sensitivity to EMP

The sensitivity of a system to disruption due to EMP is a function of the system characteristics (i.e., digital or analog, frequency of operation, configuration, etc.), and of the components used in the system.

Laboratory tests have demonstrated that vacuum tube systems are many times more resistant to permanent damage from EMP than semiconductor systems, and 60 Hz motors even more resistant than vacuum tubes; but even motors can be damaged if connected to a very large energy-collecting structure.

The replacement of "harder" vacuum tube systems by "softer" transistors or integrated circuit systems was an aspect which was never considered during the atmospheric test period. This evolution of systems using components that require inherently less power has resulted in an increased susceptibility to EMP.



Digital computers are extremely sensitive to EMP. Here it makes little difference whether vacuum tubes or transistors are used. It can affect operation of a computer by introducing false signals or by erasing information stored in memory banks. Very small amounts of energy may perturb such things as the flight path of a missile, momentarily disrupt display devices, temporarily upset the operation of many other types of electronic gear, etc.

Tests on systems have demonstrated an "avalanche effect", wherein very small amounts of EMP can "dump" huge amounts of stored electrical energy which normally occurs within a system. This can set off an electronic/electrical "landslide" which could damage components or jam communications systems. This is somewhat analogous to a tiny spark igniting a forest fire. In fact, EMP-induced "sparks" could well ignite fuel-air vapors or detonate ammunition under the proper conditions. A well-publicized example of an "electrical avalanche" due to a small component failure was the Northeast power system blackout. This, of course, was not induced by EMP.

Most of our electronic systems are designed to be frequency selective, (i.e., operate over a narrow frequency band), and, therefore, respond differently to the EMP spectrum. As we have seen, the EMP spectrum is very broad, but it certainly is finite. Where the energy couples to the system (power line, antenna, etc.) determines the spectral content and the energy transmitted to sensitive components within the system. Therefore, all systems will not be susceptible to EMP nor will they respond in the same manner or to the same extent.

Examples

What does this mean in more specific terms? A few hypothetical system examples may be illustrative. Consider a communications complex. The most likely EMP effect would be temporary interruption of communication service. This can occur even though no permanent damage requiring parts replacement occurs. Where EMP protection was not considered, it has been shown that one EMP pulse could impair service of certain portions of the system for as long as 20 to 40 minutes. Burnout of key components is somewhat less likely, but always a possibility.

In the case of strategic systems and some tactical systems, no impairment of the system can usually be tolerated, even for a few seconds. For example, EMP pickup in the control system of an in-flight missile could perturb the flight so much as to cause breakup of the missile, could disrupt the count-down of missiles being readied for firing, or could impair the command and control during critical times.

In the case of tactical systems, many of these systems are required to survive the effects of the nearby detonations of a low-yield nuclear weapon. Here it makes little sense to invest millions of dollars in blast, thermal, and TREE protection without providing a comparable degree of EMP protection as well.

Consider a hypothetical, anti-aircraft, fire-control, and fighter-director tactical system. One EMP event has the potential to "erase" all of the recorded data supplied by over-the-horizon picket aircraft and ships or distant sensors even though no permanent damage occurs. This one-to-ten minute potential service interruption could allow an enemy bomber enough of an advantage to destroy the system.

Nuclear war without a direct USA involvement is a relevant part of EMP weapon effects scenarios. A nuclear detonation, a continent away, must not functionally damage vital radio communication links from the USA to tactical units -- ships, aircraft and ground units stationed near the conflict.

The photograph dramatically illustrates some of the pseudo-focusing or field-gathering effects. In this case, the EMP field intensity has been "gathered" (or more correctly, enhanced) by a factor of about 100 times. This "pseudo-focusing" was sufficient to ionize the air near the blade tips, as shown in this photograph taken of a rotor blade of a helicoper undergoing nighttime EMP simulation tests.

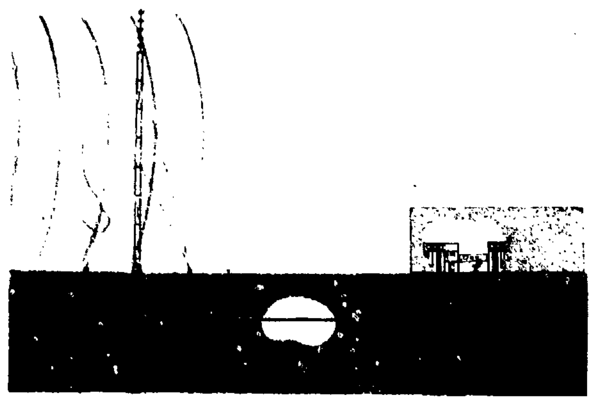
Summary

The 3,000 mile high altitude electromagnetic EMP effects radii occurs because of:

1. large source region,
2. very intense fields with characteristics never normally encountered,
3. "focusing effect" of energy collection,
4. extreme sensitivity of modern components and subsystems,
5. "house-of-cards" design of systems.

From the foregoing, we can see that EMP susceptibility is a three-factor function:

- Energy collection; collection efficienty (size) of structure,
- Fraction of collected energy applied to sensitive component,
- Sensitivity of component to damage or upset (interruption).



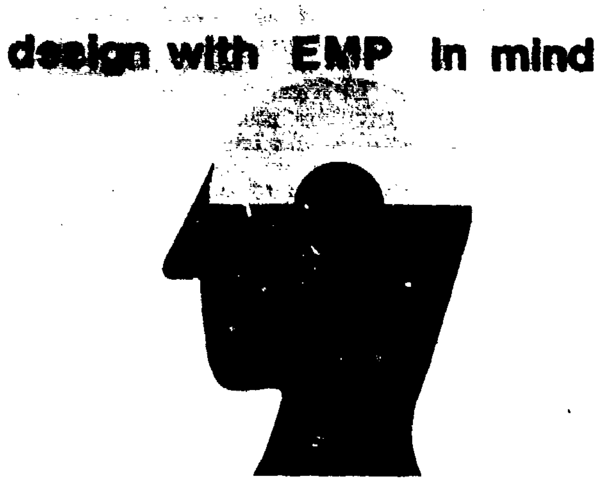
The potential disruption effects of EMP must be considered for all systems because of the ever-increasing dependence on modern sophisticated electronic/electrical systems for both military and civilian uses.

Thus all systems of any importance should incorporate features and techniques to counter the EMP effects, as required. A number of systems have already been demonstrated to be protected and many more systems are currently being EMP-hardened.

Thus, the central issue today is PROTECTION.

1.4 EMP PROTECTION PRACTICE

EMP protection is achieved through three basic activities: design and analysis, protection incorporation, and test verification. Each of these activities continues throughout the various phases of a hardened system design. Any EMP protection scheme, however well designed initially, must be implemented and continued through good quality control during production, good maintenance practices during its operational period, and control of any retrofit or field modifications.



Design and Analysis

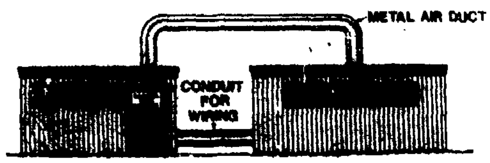
Design and analysis of protection measures is required to:

- Identify protection requirements -- define how much protection is required, and which portions of the system/subsystem/equipment needs protection.
- Select and specify the protection scheme -- how will the required protection be achieved.
- Guide the test and verification phases of the program -- establish the test requirements and level of tests (i.e., component or subsystems laboratory tests or full system tests).

Many analytical tools are available, but they are generally restricted to evaluation of idealized configurations. In order to apply these tools to real systems, it is necessary in most cases, to make simplifying assumptions. This is achieved by deriving a simplified physical model which can then be modeled mathematically for computer aided analysis.

Depicted is an example of what is perhaps one of the most common forms of EMP system coupling -- the inadvertent conductive loop.

ANALYSIS: Consider this complex as a loop antenna.



A highly simplified physical model of this structure would be a loop antenna of equivalent cross sectional area. The coupling to such a loop can be expressed mathematically and solved for the current and voltage in the loop.

Such idealizations are often adequate, because of the broad uncertainties in nonsystem factors -- such as the probable threat range the complexity of surrounding media (earth) and other installations, and the probabilities of off-nominal system employment.

More sophisticated models can also be developed, and many accurate models and techniques are available, if greater accuracy is needed. Examples of more sophisticated models and how to use them are presented in Section V.

Protection Incorporation

There are two aspects to achieving EMP protection for our systems. Protection is realized by choosing the most appropriate protection concepts and implementing these concepts through good design and hardness assurance practices.

Protection Concepts

For protection of a system against damage (that is permanent degradation), we have the alternatives of: (1) keeping the energy out of the system, (2) keeping the energy from reaching the sensitive components having allowed it to enter the system, and (3) reducing the sensitivity of the system through the use of harder components.

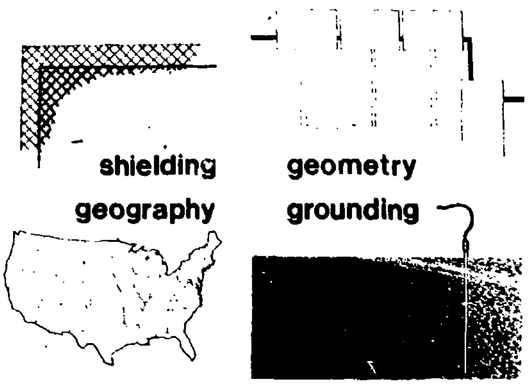
For damage protection,
REDUCE:

- EMP EXPOSURE
- COLLECTION & COUPLING EFFICIENCY
- FRACTION OF APPLIED ENERGY
- COMPONENT SUSCEPTIBILITY

Keeping the EMP energy out of the system can be achieved by reducing the EMP exposure of parts of or the entire system, or by reducing the collection and coupling efficiency of the system.

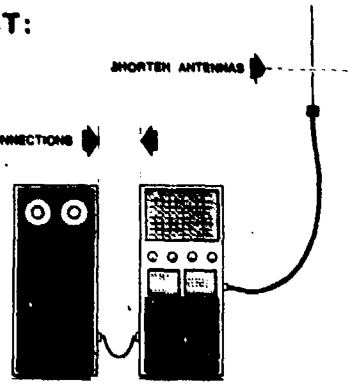
The level of the EMP environment can be reduced by:

- Shielding the system, or portions of it, in metallic shields or by earth burial, for example.
- Geometric arrangement can result in fewer or shorter cable runs or different coupling geometry.
- Geographic location, especially if multiple redundant systems are employed, can reduce the environment at least for some of the systems for surface bursts.
- Grounding control to minimize inadvertent coupling loops.



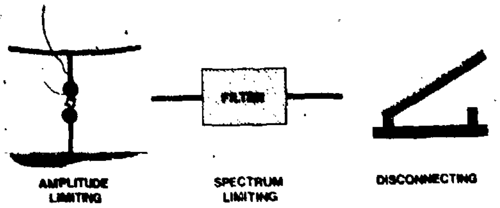
The coupling efficiency of the energy collector may be reduced by reducing its effective collection area. For example, one may shorten an antenna, for certain HF band or strong signal conditions, or rearrange equipment in order to shorten cable runs between units. Nonconducting data transmission systems, such as fiber optic or dielectric wave-guide systems, are also a good alternative to eliminate the cable coupling problem.

TO PROTECT:

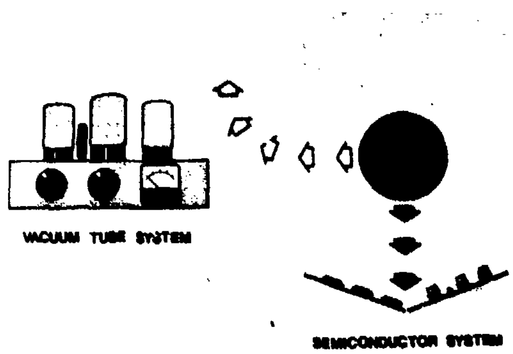


The fraction of the collected energy which is applied to a sensitive component may be reduced by introducing an amplitude limiting device, such as a spark-gap, a filter, or a disconnect mechanism between the energy collector and the component.

Reduce the applied energy by:



Protection against EMP may also be achieved by choosing less sensitive components or subsystems. For instance, vacuum tubes are more damage-resistant than transistors. Their incorporation, however, might create other problems. Semiconductors that are inherently more resistant to transient damage could, however, be selected.

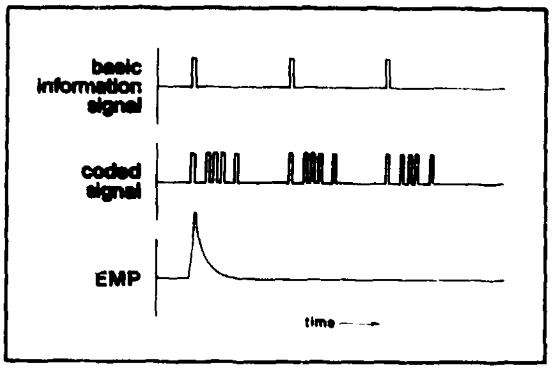


In addition to permanent degradation (damage), there is the problem of operational upset in many systems. Operational upset can occur at much lower levels than damage, and therefore requires a greater degree of protection. All of the concepts associated with protection against permanent damage can also be applied for upset protection. There are additional hardware and software approaches available to minimize the upset problem.

CONCEPTS FOR UPSET PROTECTION

- ALL DAMAGE PROTECTION MEASURES
- HIGH LEVEL DIGITAL LOGIC
- CODING
- HARD MEMORIES
- EMP EVENT SENSING
- SOFTWARE CIRCUMVENTION

Using high-level digital logic, that is logic circuits which have higher voltage switching thresholds, reduces the probability that the EMP coupled pulse will be of sufficient amplitude to result in a logic state change. The use of reasonably long coded signal chains can further reduce the probability that the logic would err in terms of recognizing the EMP induced signal as the correct signal.



As in the case of damage protection, the use of harder components can help to minimize the upset problem. The use of less easily upset memory systems, such as magnetic drum or disk memories, is one technique. This would permit storage of important segments of processed information or portions of the program in a memory which would preserve this information.

Error sensing and correction codes could be included in the software to further reduce the probability of upset or errors in the output data. In some cases, this is the only hardening that need be implemented.

EMP event sensing can be used for error correction or to inform system operators that EMP may have introduced an error. In some instances, this could be achieved automatically through the computer software, such as an instruction to reprocess the information starting at a point prior to receipt of the event signal.

Design Practices of Protection Concepts

The foregoing are only protection concepts which must be properly implemented. The applicability of these concepts to any system depends on the system requirements including both operational performance and environmental factors. The design practices are the practical details associated with implementing the concepts.

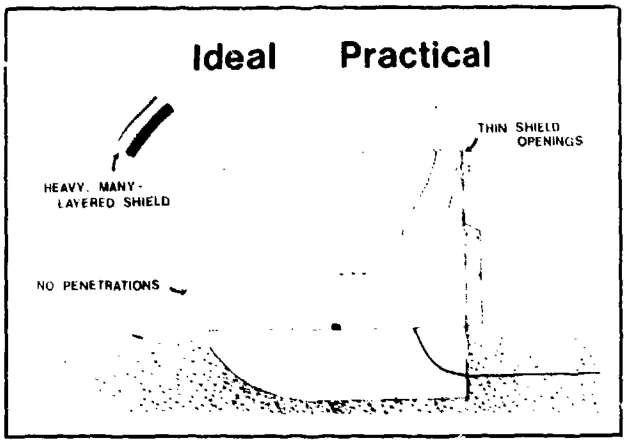
There are several viewpoints toward a rational, balanced, and complete consideration of these "design practices". The approach taken in this course is to categorize these practices according to a hierarchy of application of the practice ranging from the system level (overall system protection) to terminal protection using protective devices.

CATEGORIES OF PROTECTIVE PRACTICES

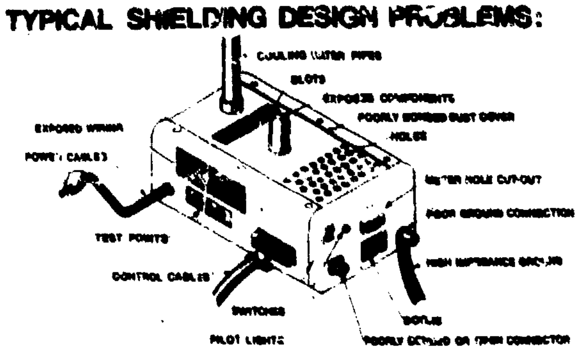
- SYSTEM ASPECTS
- SHIELDING
- CIRCUIT LAYOUT
- GROUNDING
- CABLING
- PROTECTIVE DEVICES

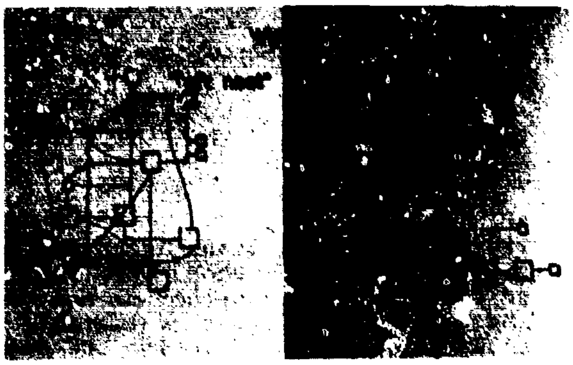
In this overview section, we will look at a single feature of each of these for orientation purposes.

The systems aspects are concerned primarily with the concept of keeping the undesired energy (in this case EMP) out of the system. As such, these aspects deal with the control of departures from an ideal system from an EMP viewpoint. The ideal EMP protected system is depicted in the left hand portion of the figure. "Reality" usually involves a large number of potential "violations" in terms of apertures, conductor penetrations, and so forth, as depicted on the right.



The circuit layout aspects are closely akin to the systems aspects. At the system level they deal with the equipment and intercabling configuration. At the other extreme, the circuit card, they deal with layout of printed circuit boards. The dominant principle is the avoidance of coupling configurations -- most notably inductive loops. Good circuit layout practices apply equally to large cable systems or to printed circuit packages. It is generally independant of circuit dimensions.





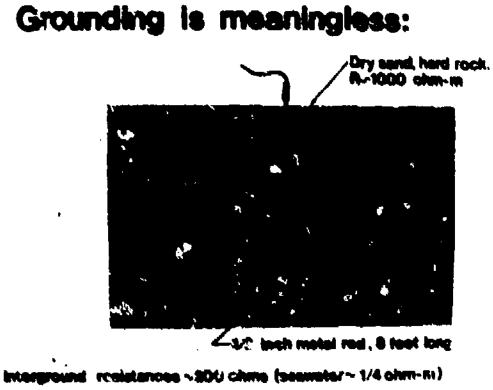
Grounding is not a panacea, nor should it be viewed as an EMP protection practice by itself. However, if it is not realistically viewed, it __may__ make things worse. In some nuclear test instrumentation systems, __controlled resistive__ grounds are purposely used to promote energy dissipation and to suppress ground loop currents.

Here we see the most common -- and most useless -- "ground".

Shielding design usually centers on pragmatic compromises related to the realities of construction and fabrication, economics, and applicational requirements. These generally boil down to:

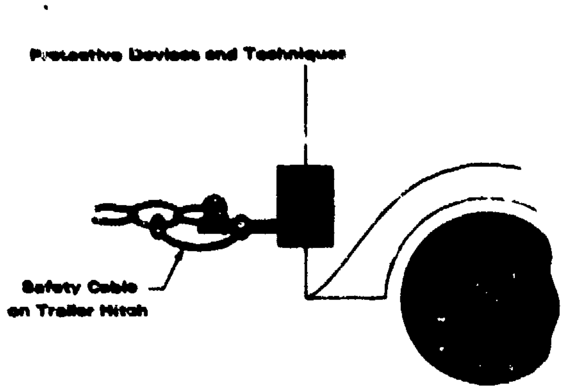
- Wall thickness and integrity
- Apertures -- "tightness"
- Penetrations -- necessary conductors.

Good shielding design criteria apply to all levels, that is overall system envelope shields down to shielding of individual circuits.

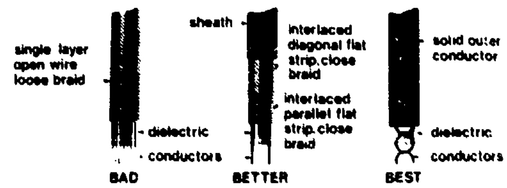


Cabling

Cable design represents an extension of shielding and circuit practices in the EMI viewpoint. It is the area in which the worst compromises are often made in the interests of economy. Good cabling can be expensive, but EMP coupling via bad cabling can be fatal. Here we illustrate this trend in cable construction. (There may be additional shields around individual internal conductors, of course, which are not shown here). As stated previously, nonconducting data transmission links offer a potential alternative to the use of shielded cables. Their applicability must be determined based on the system performance requirements, reliability of fiber optic or dielectric wave-guide systems, cost, etc.



CABLING-THE "OUTER CONDUCTOR"

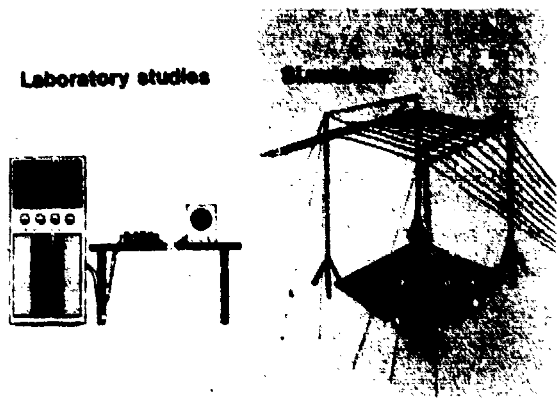


Testing

Protection verification is essential. Originally, such test methods were addressed to determining whether or not existing systems really were vulnerable. Today, testing is used to verify the design hardness of systems or equipments.

As important as planning and implementation of hardware protection is its testing. The balances between laboratory tests and full-scale simulations, and between component and full-system tests, depends on many factors, such as system size, probable threat situations, unavoidable susceptibilities, and so forth.

Protective devices such as surge arresters, filters, and circuit breakers are generally used after the other aspects have been applied. It is not possible to keep all the coupled energy out of the system. These terminal protection devices are to further limit this residual energy at the terminals of equipments employing sensitive components or circuits from either damage or upset.



The reasons for testing are summarized below:

Why Test?

- Verification of analysis
- Extention of analysis
- Surprises
- Quality assurance
- Certification

Testing, in some cases, is for the purpose of verifying analysis and the design of our systems. It provides confidence that what we did was correct. On the other hand, due to the required simplifying assumptions used in the analysis phase, the test phase reveals obscure coupling paths or system degradation (surprises). Finally, testing provides us a means of assuring the quality of the end product and for production control.

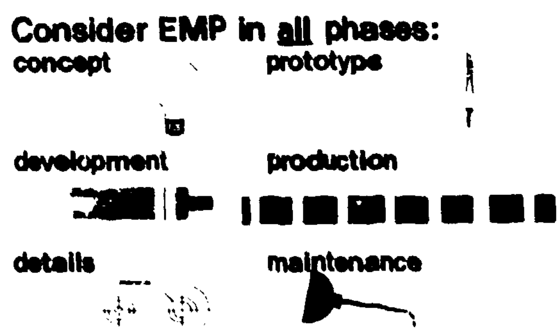
To illustrate the need for testing, several typical surge protection components were tested. EMP waveforms, having very fast rise times, short durations, and high peak voltage, were applied instead of the usual test waveforms based on lightning or other more common surge-test requirements. In the cases of two of the protection approaches, the data sheet ratings were not exceeded by the EMP test waveforms. In all cases, these protection components were destroyed. High current, fast rise-time pulses can be applied directly to the circuit being tested, using a high-power pulse generator. Such lab tests can add assurance that the system will pass full-scale simulation tests.

A simulation of the radiated EMP environment can also be created (see Section VII). For example, the ARES simulator is built as a very large transmission line and produces EMP-like fields in the space between the towers.

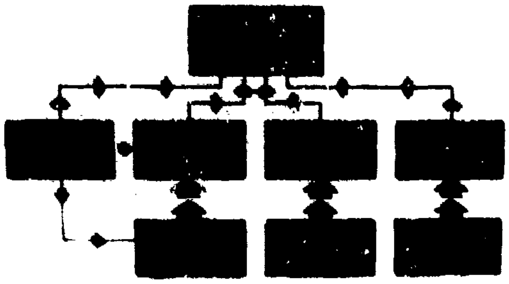
A final test of the complete system, in a full-scale simulator, is a desirable goal. But, at present, this is sometimes impractical, for example, in an extended communication network or free-field illumination that has to be augmented with direct pulse injection techniques.

1.5 PROGRAM PLANNING

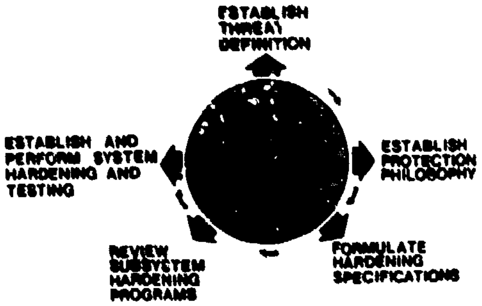
EMP is a system problem and, therefore, must be considered on a system basis during all phases of system design.



The specific management approaches to EMP for any given system will depend on many factors -- people, funds, equipments, existing organizational structure -- but must be considered at all levels. Management must have visibility of the EMP control program and provide authoritative direction for resolving conflicting requirements.



EMP considerations at the system level are concerned initially with threat definitions, and establishing the protection philosophy. As the program progresses, management is concerned with review of the hardening program at the system and subsystem levels, and insuring that the hardening is properly incorporated.



EMP is a challenging management problem. Uncertainties regarding analysis and testing are the basis for this challenge. More important, the state-of-the-art of EMP is still evolving. Time and funding allowances must be made for adequate personnel training. Lastly, as reiterated previously, EMP must be considered for the entire life of the system.

EMP IS A CHALLENGING MANAGEMENT PROBLEM

- Uncertainties of Analyses
- Difficulties of Testing
- State-of-Art
- Limited Facilities
- Personnel Training
- Total System Life

Part of the challenge is the state-of-the-art. The currently available information, as exemplified by the Handbooks and other references in the area, is directed primarily to a discussion of the EMP phenomenology, electromagnetic interaction, hardening concepts and implementation guides.

The more developed electromagnetic interaction areas, such as radio frequency interference or lightning, have available commonly accepted design nomographs and charts which can be used reliably to select components or implement designs.

This "technology base" for RFI and lightning has taken many years and millions of dollars to develop. A similar development has yet to "mature" for the EMP problem area. As a result, the currently available information must be translated into the specific needs of the EMP hardening plan. This will take money, people, and training time for this purpose. In other words, there is no easy way out at the present time.

Currently Available Information

Phenomenology   Coupling   Response   Hardening Concepts   Examples
In
Interaction   Implementation Guides

\$ + People + Training Time

Translated

Fixes, Nomographs, Charts, Components, Hardening Plan, Test Plan, Maintenance Plan, Construction Guide, How-To-Do-It Cook Books, Specifications, Test Procedures, Quality Control, Modification Procedures, Field Personnel Training, Retrofit Control

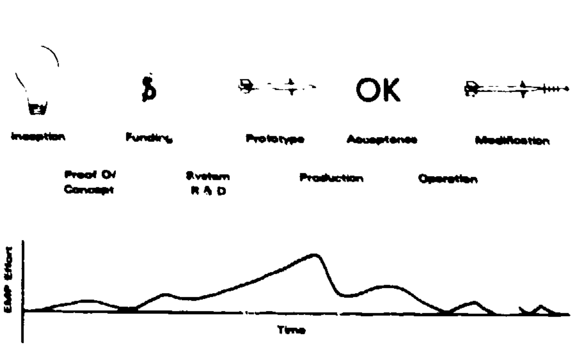
An adequate base of personnel qualified to handle EMP problems often does not come easily. Management should consider a planned education period for the training and education of personnel. Past experience has indicated at least one to two years of direct involvement is required to develop a capability for adequately dealing with EMP interaction and hardening problems.

In addition, this training period should allow for a few mistakes. One such approach is to initiate the EMP hardening efforts with what might be termed a "test-bed" program. During the course of this test-bed program, the necessary analysis, testing, and hardening are developed in terms of a consistently funded and well-laid out effort.

PERSONNEL CHALLENGES

- Planned Capabilities Development
- One To Two Years
- Plan For A Few Mistakes

Hardening a system against EMP is not a one-time effort. EMP protection must be considered from the time of inception throughout its life cycle. If the system is to remain hard, EMP control must be exercised during the production and operational phases including control of all modifications.



One of the major challenges to management is meeting the funding requirements to provide adequate hardening. In many cases, this can be done during the initial program planning phases. If this is not always possible, especially for a program in being, one approach has been the so-called hardware trade-off approach, which trades off a large number of systems of unknown survivability for a somewhat smaller number of systems with assured survivability.

Another approach has been to minimize the need for hardening for the close-in source-region threats by employing geographic dispersal and increasing the number of systems somewhat.

Another approach can consider minimizing peacetime reliability costs and converting the dollar savings into assured survivability.

MEETING THE $ CHALLENGE

Program $ Initially

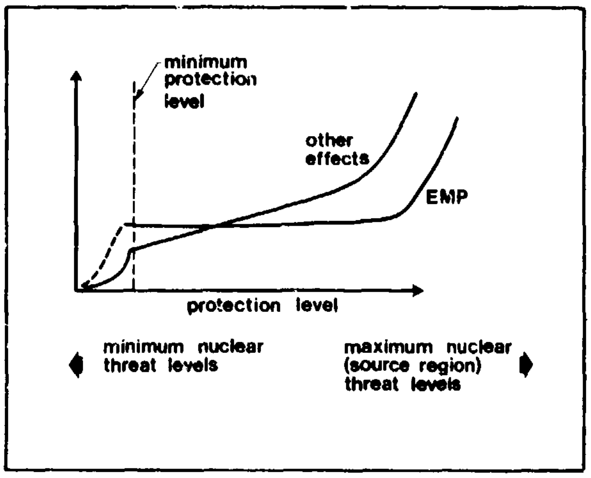
OR

Trade-Off Large Number Systems of Unknown

FOR

Somewhat Smaller Number of Systems With Assured Survivability.

In general, the bulk of EMP hardening costs are roughly independent of the required protection level, once the decision is made to protect. Typical EMP hardening costs (if incorporated early) for strategic systems can be on the order of a few percent of the system cost. If retrofitted, costs can easily rise in excess of 10 percent of system cost, to nearly the original system cost in some cases.



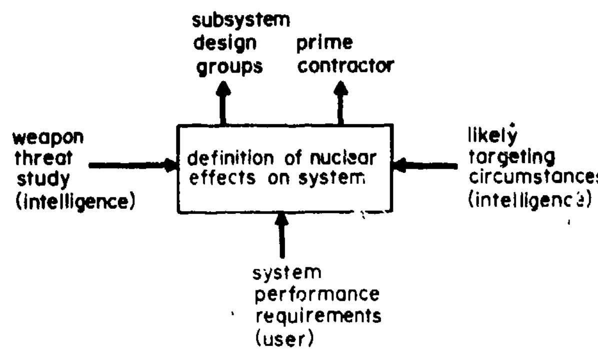
Management must also provide for the necessary equipments tc conduct the EMP hardening program. Typical equipment requirements for a complete EMP hardening program would range from computers, laboratory impulse test equipment to large-scale simulators.

Contractors (both prime and, in some cases, subcontractors) could be expected to possess or have access to the necessary computers to conduct the required analyses. Depending on the complexity of the test requirements, contractors could be expected to have at least limited laboratory test setups. For the more complex subsystem type tests, sophisticated laboratory and field testers are available at the lead laboratories of the Army, Navy and Air Force. These large-scale free-field simulation facilities are constructed, maintained and operated by DNA, and the three services' lead laboratories.

In order for management to establish, review and control an EMP program, especially for large complex systems, they must have a formalized reporting structure. These reports have often been denoted as "White Paper" and assessment reports.

Identifying the nuclear weapon threats to the system in context of the mission of the system is often called a "White Paper".

Delineation of these threats determines the resulting EMP environments which must be considered.



The assessment reports initially translate the EMP threats in terms of the testing, analysis and hardening requirements. Later on they can be used to review progress toward achieving an EMP-hardened system.

ASSESSMENT REPORTS

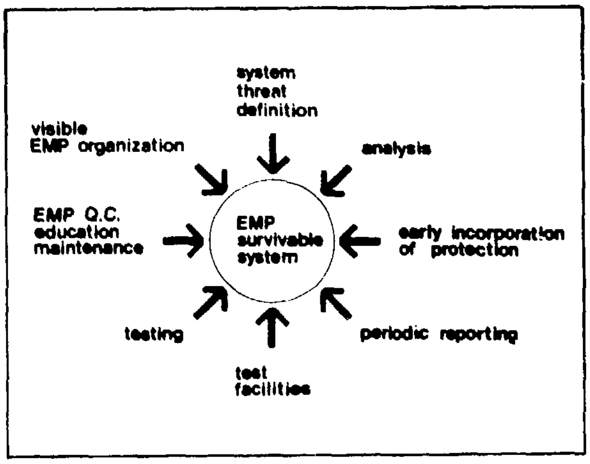
PERIODICALLY INDICATE STATUS IN:

- HARDENING CRITERIA AND IMPLEMENTATION
- TESTING
- ANALYSIS

On the basis of past experience, it is suggested that careful review of the EMP program be periodically conducted for possible trouble areas, such as preoccupation with one area, for example, shielding or terminal protection, at the expense of other areas. Above all, the EMP program must have good visibility and management.

The EMP hardening features must be maintained during production by proper quality control and during use by education and training of personnel, as well as by periodic inspection and retesting.

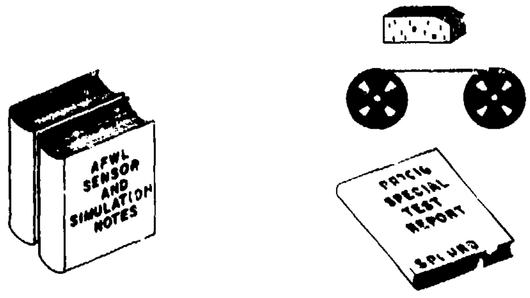
While many of the program planning aspects are self-evident, the most important are summarized here.



1.6 STATE-OF-THE-ART

Today, the state-of-the-art is considered to be sufficiently advanced to permit the cost effective hardening of most military hardware with a high assurance of acceptable EMP hardness level. The information that is available is in a form to provide guidance in the design and implementation. The answers to specific questions are not generally available if programs have not been in existence to attack them.

A number of information sources are available, such as the DNA handbook, service laboratory handbooks, DNA Protection Engineering and Management Notes, AFWL Note series, technical reports, specific system reports, and so forth.



EMP System Vulnerability   Lead Lab Consultation

A library is maintained by DNA-DASIAC at G.E. Tempo in Santa Barbara, California which includes most of the report documentation. A computer program library is maintained by the Lawrence Livermore Laboratory, Livermore, California under DNA funding. Data files (computer storage and retrieval) on semiconductor and component damage levels, computer programs of special nature, and technical libraries are also maintained at the service lead laboratories.

Both analytical and empirical approaches are required to realize a hardened system. Neither can be exclusively relied upon. The number of uncertainties and unknowns is so large that one approach is often used to confirm the results of the other approach.

LIMITS of ANALYSIS

LIMITS of TESTING

At the present time, our analytical capability is sufficient to assess the EMP hardness of systems and to design future systems EMP hard with reasonable confidence. There are still limitations, due to the assumptions we must make, to our analytical capability so that we must still reply on testing to verify the analysis and uncover obscure problems.

The specific EMP response is highly dependent on obscure details not generally controlled during manufacture or design (such as the type of corrosion protection employed, extraneous coupling path, etc.) Further, the number of details important in the electromagnetic interaction sense can be quite large. For example, typical electronics subsystems may have as many as ten thousand potentially susceptible electronics components and related coupling paths. The electrodynamic analytical techniques are amenable only to fairly simple structures and limited number of components. Specific component responses cannot be analytically modelled because much of the basic information has yet to be developed.

ANALYSES AND STUDIES

CAN:
- find problem areas
- identify likely areas of weakness
- provide design assurance
- guide test approach selections
- confirm test results

CANNOT:
- identify specific levels of vulnerability or susceptibility for a complex system
- find specific weak points
- select more or less susceptible components

WITHOUT RESORTING TO TESTING

Like other electromagnetic problems in RFI (Radio Frequency Interference), HERO (Hazardous Electromagnetic Radiation Effects on Ordnance), or TEMPEST (Compromising Electromagnetic Emissions from Secure Communication Equipment), empirical testing is required. However, an exclusively empirical approach has obvious limitations because the very act of making a measurement can upset the test and introduce extraneous data. Further, testing alone can become very expensive if all aspects of system hardness must be developed experimentally, especially for widely dispersed systems or for systems which must survive above ground within the source regions. Finally, many times a test specimen is not available, such as during the design phase, so analysis is the only available tool.

TESTING

CAN:
- validate analytical approaches
- confirm design
- develop component subsystem susceptibility
- reveal obscure electromagnetic details
- provide final quality assurance

for compact system

CANNOT:
- relate test method to threat
- determine if tests are properly conducted
- economically test widely dispersed systems
- test for many "source region" threats

WITHOUT RESORTING TO ANALYSES AND STUDIES

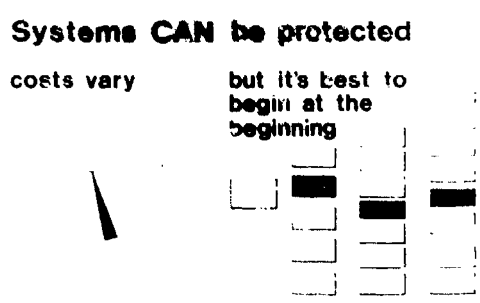
1.7 SUMMARY

In summary, the EMP from a high-altitude burst can affect substantial fractions of the earth's surface without significant contributions from other weapons effects. The waveshape, field intensity and spectral content of the pulse are unlike those normally experienced due to man-made or other natural sources.

Hardening to EMP from a high-altitude burst must be considered for almost all important systems, even though other weapons effects may not be important.

The EMP from a surface burst is more restricted spatially. It has an exceptionally large and abnormal waveshape within the source region. Within the source region (or near to it) where the EMP poses a major threat, the other nuclear weapons affects must also be considered. The system must possess "Balanced Hardness" with respect to the other nuclear weapons effects of blast, thermal, radiation, etc.

Based on somewhat scattered data developed during the period of atmospheric testing and extensive information based on thorough analysis and experimental simulation, EMP can cause functional damage or operational upset, especially for the more sophisticated systems employing transistors or those highly dependent on digital computation. Today electrical and electronic systems can be protected. The cost of this protection will vary with the nature and mission of the system. The most economical and effective protection is realized if the hardening effort is considered early and made an integral part of the system design.



The rest of the course will provide a more detailed and technical insight into how to carry out an EMP-hardening program.

It is designed to provide an <u>awareness</u> of the various approaches which <u>might</u> be used for your system.

The various "tools," analytical or experimental, will be discussed and their advantages and limitations presented.

A few examples will be given on how these "tools" may be used.

But we cannot make "experts" overnight, nor will attending this course provide all of the answers you need.

THE ENGINEERING COURSE
WHICH FOLLOWS WILL PRESENT:

- WHAT "TOOLS" ARE AVAILABLE
- WHEN THESE MIGHT BE USED
- HOW TO USE THEM

SECTION II

INTRODUCTION TO THE TECHNICAL COURSE

As an introduction to the technical portion of the course, the objectives of the course and the content of the course will be stated. Further, some basic electromagnetic principles, which are important for understanding what will be covered subsequently, will be reviewed. Finally, some EMP and system characteristics which are referred to throughout the course will be defined and discussed.

2.1 COURSE OBJECTIVES

The overall objectives of this course are to provide an awareness of what EMP is, what its impact on systems may be, what tools are available to assess the impact on systems, and how to protect against the effects of EMP.

Many of the answers are still being worked out for particular systems applications in regard to the:

Methodologies

Analyses

Protective Techniques

Test Techniques

Maintenance

These answers are in a state of evolution.

EMP Hardening is a Difficult Problem	
Questions	**Answers**
Methodologies	?
Analyses	?
Protective Techniques	?
Test Techniques	?
Maintenance	?

As a consequence:

- There are no easy, optimum, or cookbook approaches which are applicable to all systems. Today, each system is treated as an independent entity, using past efforts for guidance.
- There are no universal hardening techniques or devices which apply to all cases. The required protection for each case must be determined based on the threat and the mission requirements.
- There are no optimum analytical or test approaches for all cases. These must be determined based on the system configuration, complexity, goals of the program, and costs.

Hence, we will present to you several approaches to the EMP problem. We hope to provide enough background to enable you to select the "best" combination for your needs.

There are No Easy Answers

No "One and Only"

Right-Way

The course can only reflect the state-of-the-art; so do not expect to come away from the course loaded with "nomograph-chart" solutions, "cookbook" formulas or standardized approaches which can be briefly reviewed and routinely applied to your system.

We hope to tell you something about the problem, what "tools" are available to help solve the problem, the limitation and advantages of "the tools," what the hardening concepts are and how to apply them, and how some people are approaching the overall problem. We hope this will be sufficient for a first-cut assessment planning effort to formulate programs for dealing with the EMP problem. Examples to illustrate these will also be given.

What to Expect?

The State-of-the-Art.

* Concepts
* Methodologies
* Tools
* Applications
* Examples

In three days we cannot turn you into EMP specialists.

We do not intend to teach you basic mathematical processes, such as Fourier transformations, how to program SCEPTRE, or other computer codes, etc.

We are not going to make you into a nuclear physicist, simulator designer, or high-voltage specialist.

You yourselves will have to apply and enlarge your existing skills to the EMP problem based to some extent upon the summary of all the disciplines, skills and approaches presented during the course.

In the three days allocated to the "EMP" Awareness Course," we can only hit the high points of a multi-disciplined problem. References are provided to assist you in your future efforts and applying the information and tools which have been developed.

In Three Days We Can Only Hit the High Points of a Multi-Disciplined Problem	
Nuclear Physics	Program Management
Applied Math	Design Practices
Field Theory	Quality Control
Circuit Analysis	EM Maintenance
Field Measurements	Operational Training
Test Simulation	

2.2 BASIC ELECTROMAGNETIC PRINCIPLES

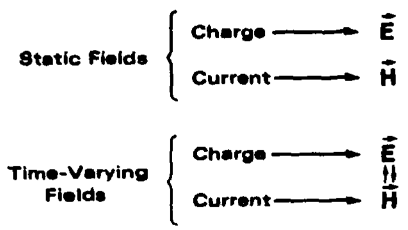
To adequately describe or characterize electromagnetic phenomena in nature (in this case the EMP phenomenon), requires familiarity and understanding of some basic field quantities. These field quantities are listed below along with their units in the rationalized MKS system of units as a review.

ELECTROMAGNETIC FIELD QUANTITIES	
ELECTRIC	MAGNETIC
$\vec{E}$ = Electric Field Intensity (volts/m)	$\vec{H}$ = Magnetic Field Intensity (amps/m)
Ψ = Electric Flux (coulombs)	Φ = Magnetic Flux (webers)
$\vec{D}$ = Electric Flux Intensity (coulombs/m^2)	$\vec{B}$ = Magnetic Flux Density (webers/m^2)

It is essential to keep in mind that in a time varying electromagnetic field, the electric field ($\vec{E}$) and the magnetic field ($\vec{H}$) cannot be created independently. Coexistence of the electric and magnetic fields is a prerequisite to the establishment of an electromagnetic field. At low frequencies ($\lesssim$ 10 kHz), the electric and magnetic fields are often considered separately for simplification of shielding analysis and design. At these low frequencies this is a very good approximation. True separation of these fields only exists in the static (dc) ase, however.

Parameters of the Medium	Relationships Between Field Quantities and Parameters of the Medium
ϵ = Permittivity (Farad/m)	$\vec{D} = \epsilon\vec{E}$
μ = Permeability (Henrys/m)	$\vec{B} = \mu\vec{H}$
σ = Conductivity (Mhos/m)	$\vec{J} = \sigma\vec{E}$

Coupling of Time-Varying Electric and Magnetic Fields



The relationship between the electric and magnetic field is given by:

$$\frac{\vec{E}}{\vec{H}} = \eta$$

where

η = characteristic impedance of the media

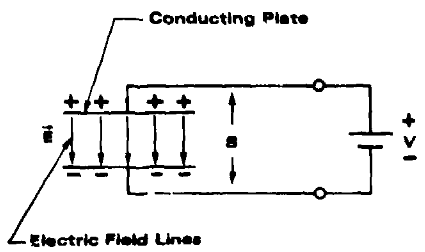
η_o = 377 Ω is the characteristic impedance of free space.

The medium also plays an important role in formulating the electromagnetic field. The parameters of the medium provide the necessary link between various electromagnetic quantities. The more important medium parameters and their units, and the role they play are indicated here.

Electric fields are the result of charge separation in the media. For simplicity, consider the generation of a static (dc) electric field. A simple case is that of a parallel plate capacitor. Impressing a voltage (V) across the plates results in a redistribution of charge on the plates as shown. This redistribution of charge results in an electric field ($\vec{E}$) between the plates. The relation between the electric field and the applied voltage, in this case, is given by:

$$\vec{E} = \frac{V}{S} \text{ volts/meter}$$

Generation of an Electric Field in a Parallel Plate Capacitor



If we have a force which causes charge separation, an electric field is created. In this case, the force was the applied voltage. For example, in the case of EMP generation, as we will see in Section III, the force causing the charge separation is the nuclear detonation and the resulting gamma rays.

The generation of magnetic fields requires current to flow. Again for simplicity, consider the static (dc) case. Consider, for a simple example, the field between two current sheets carrying a total current I. The total current is related to the current density (J_s) by the relation

$$I = J_s S$$

where

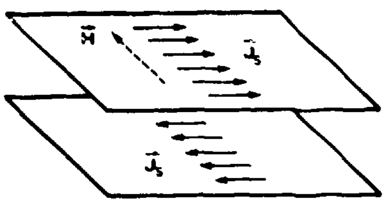
S = cross sectional area of the current sheet.

This assumes a uniform current density which holds only for the static case. The resulting magnetic field ($\vec{H}$) between the current sheets is given by:

$$\vec{H} = \frac{\vec{J}_s}{2\pi d} \text{ amps/meter.}$$

In the case of EMP, see Section III, the current flow results from the interaction of the gammas produced by the nuclear detonation with the media.

Generation of a Magnetic Field Between Current Sheets



In many cases in field theory, the events occurring at a point, or within an incremental volume, are of interest. In these circumstances, it is desirable to define the relationships between the current and voltage for the incremental volume. These are shown in the figure. The concept of current density is useful. For an incremental volume or area, the current is uniform throughout, and the current is related to the current density by:

$$I = \vec{J}S = \frac{V}{R}$$

where

$\vec{J} = \sigma \vec{E}$

$V = \vec{E}L$

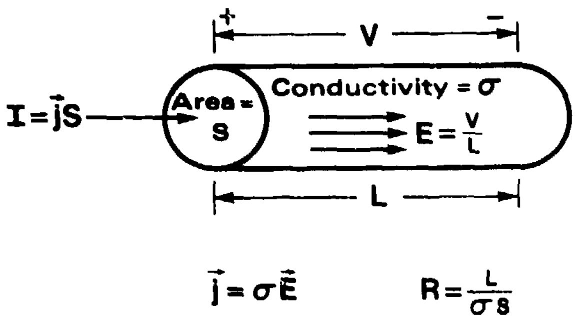
$R = \frac{L}{\sigma S}$

σ = conductivity of the material

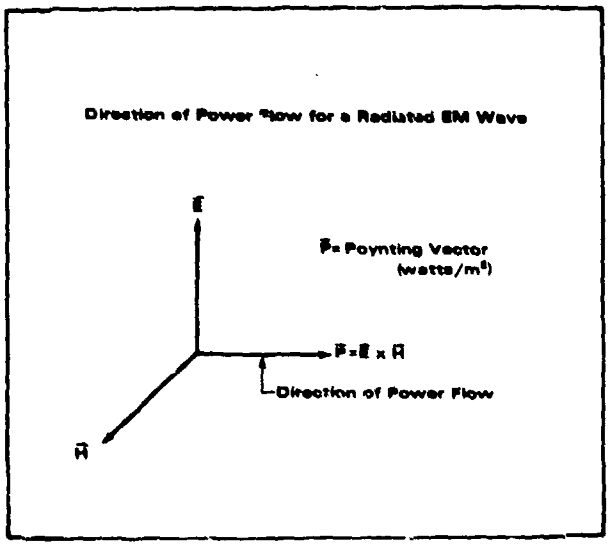
R = resistance of the material for the incremental volume

$\vec{E}$ = induced electric field.

Ohm's Law for an Elemental Volume



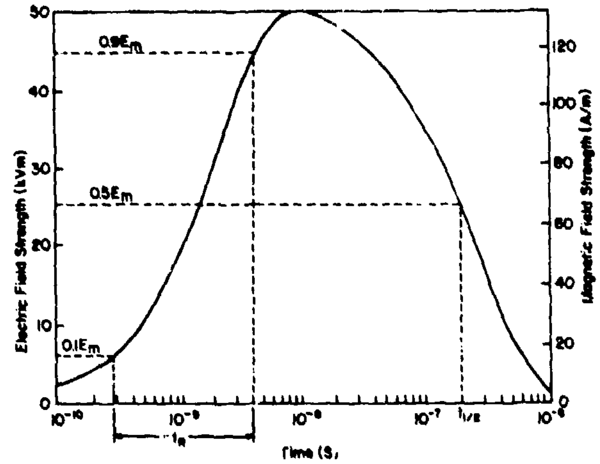
Direction of power flow in a distant EM field is given by the POYNTING vector and can be obtained by use of the right-hand rule, as indicated. Note that the POYNTING vector, $\vec{P}$, gives not only the direction of power flow but also the power density of an EM field in watts/m^2.



2.3 EMP CHARACTERISTICS

The waveform of the EMP plays an important role in the coupling to systems and the effects on systems. At this time we will define the more important characteristics of the EMP to serve as a base of understanding for subsequent discussions.

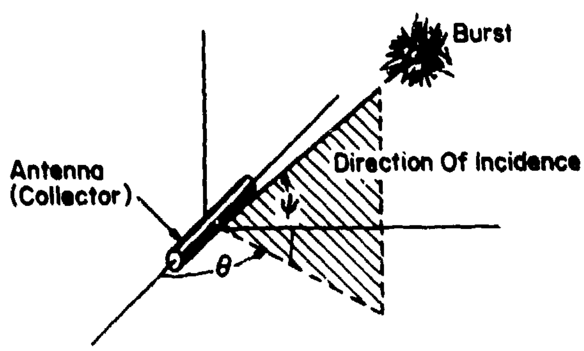
Two of the most important parameters are shown in the figure. The rise time (t_r) is defined as the time between the 10 and 90 percent points of the leading edge of the wave. The time to half-value is the time for the wave to decay to half amplitude on the trailing edge of the pulse. These times determine the spectral content of the wave.



GENERALIZED HIGH-ALTITUDE EMP ELECTRIC AND MAGNETIC-FIELD TIME WAVEFORM

Equally important, in order to determine the coupling to a system, is the polarization of the wave. The polarization is defined on the basis of the $\vec{E}$ field vector. Vertical polarization is when the $\vec{E}$ field vector is normal to the direction or propagation and wholly within the plane of incidence. Horizontal polarization is when the $\vec{E}$ field vector is normal to both the plane of propagation and the plane of incidence.

The angles of incidence are the vertical angle (ψ) between the point of observation and the burst point measured from the location of the point of observation, and the horizontal angle (θ) between the axis of the energy collector and the direction plane of incidence.



ANGLES OF INCIDENCE

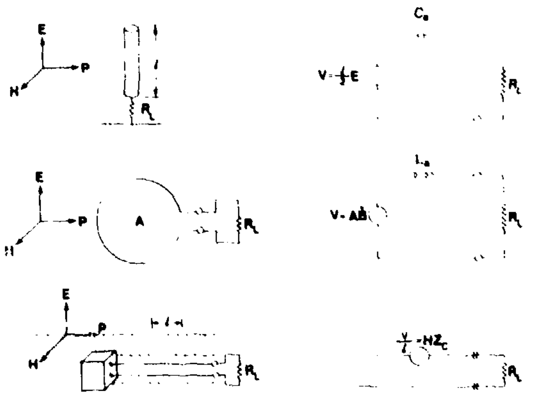
Other important parameters of the EMP are the power density w/m^2) which gives the total energy contained in the pulse, the power flow which was defined previously for an EM wave (i.e., the POYNTING vector), and the relationship between the $\vec{E}$ and $\vec{H}$ fields (i.e., the intrinsic impedance of the wave $\eta = \vec{E}/\vec{H}$).

2.4 SYSTEM/EQUIPMENT CHARACTERISTICS

The characteristics of a system/subsystem/equipment from an EMP viewpoint are those that influence (1) how the energy enters the system, (2) how much energy enters the system, and (3) what effect this energy has on the system.

Insofar as EMP coupling is concerned, energy collection by the system can be related to $\vec{E}$ field coupling (linear antennas or collectors), $\vec{H}$ field coupling (loop antennas or cable loops), or through an intermediate conversion impedance (Z_c), such as the earth, where the pickup is sometimes directly proportional to the apparent applied fields. These coupling modes apply to both exposed collectors or collectors inside shielded enclosures where the fields are reduced due to the shielding. Examples of these coupling modes are shown for the simplified case of small structures (small compared to the wavelength of the highest frequency of interest, or less than about 2 feet in length or diameter for a typical EMP waveform).

EMP Coupling

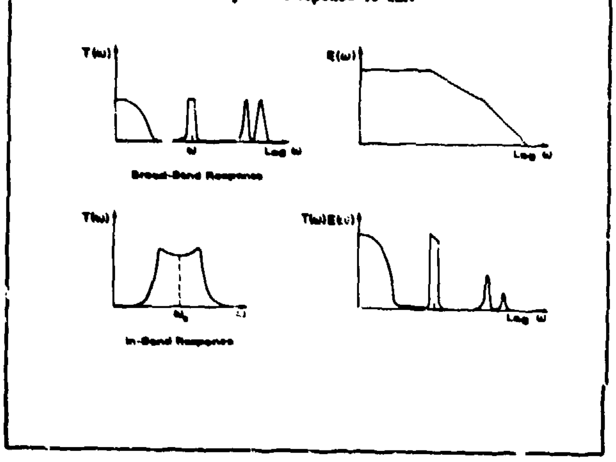


For larger structures, the general case, the collection area and efficiency must be defined. This is a function of the system size and configuration. The important aspect here is that, in general, systems cannot be considered as small structures over the very large frequency spectrum of the EMP. Consequently, to determine how much energy is coupled to the system, the response of the coupling structure or collector at all frequencies in the spectrum of the EMP must be determined.

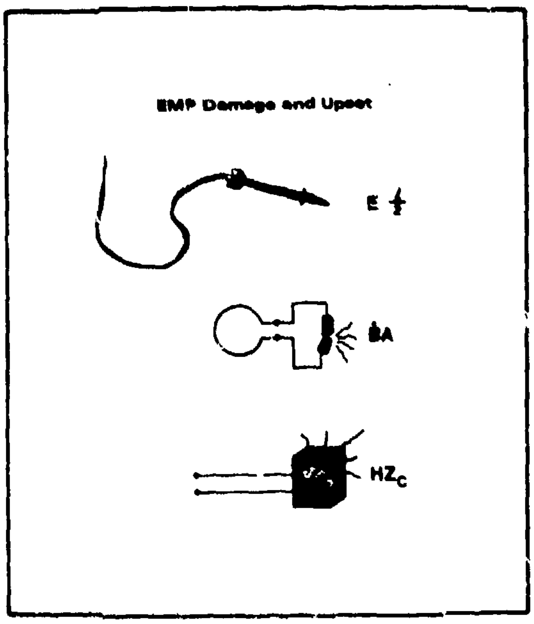
Other ways that energy enters a system are through penetrations through the enclosure. These penetrations are such things as holes in a shield (conducting enclosure), cable entries, or other conductor entries (such as water pipes, etc.).

How much energy enters the system and arrives at sensitive portions of the system is determined by the transfer function of the system. The transfer function is the ratio of the output energy versus the incident energy as a function of frequency.

System Response to EMP



Once the energy reaches a sensitive portion of the system, it may result in damage or upset. Damage is defined as permanent degradation of the system, such as component burnout. Upset, on the other hand, is defined as the temporary impairment of the system, such as circuit breakers tripping or memory erasure in computers, this results in a functional impact.



Systems are often referred to as being vulnerable to EMP. For a system to be vulnerable to EMP implies that due to its response to the EMP stimulus it can no longer perform its mission.

System hardness is another term which should be defined. The hardness level of a system is a measure of its vulnerability to the EMP. To say a system is hard implies that the presence of EMP will not adversely affect the performance of the system.

System protection (as used in Section VI) is defined as the measures (fixes) employed to render the system hard, (that is, protect the system against the effects of EMP).

2.5 SUMMARY

This section has served as an introduction to the EMP problem and presents some of the important characteristics of the EMP and systems. Further, it provides some basic knowledge of EM principles which are required for understanding the discussions.

SECTION III

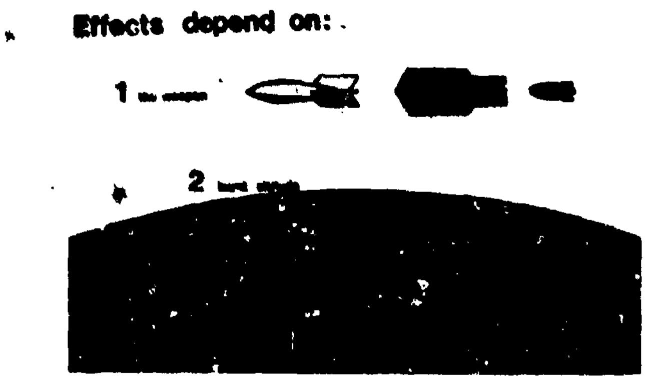
EMP GENERATION AND CHARACTERISTICS

3.1 INTRODUCTION

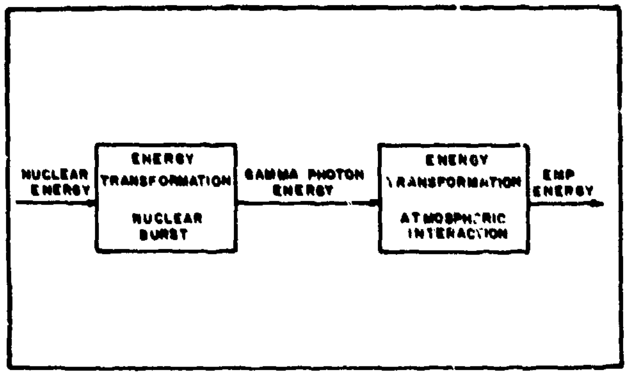
A nuclear detonation releases large amounts of energy, a portion of which is transformed into an electromagnetic pulse (EMP). The basic generation mechanism is the same for the exoatmospheric burst (height of burst greater than 40 km), the air burst (2 to 20 km), and the near surface burst (less than 2 km). The EMP characteristics and the coverage on the earth's surface are considerably different for each of the burst locations.

The relative importance of all nuclear weapons effects, EMP, blast, thermal and so forth, depends on the weapon characteristics, the burst altitude with respect to the earth, and the system position relative to the burst location. Attention will be focused on the radiated EMP produced by the exoatmospheric and near surface burst. The radiated EMP due to an air burst produces a less significant EMP problem for most systems.

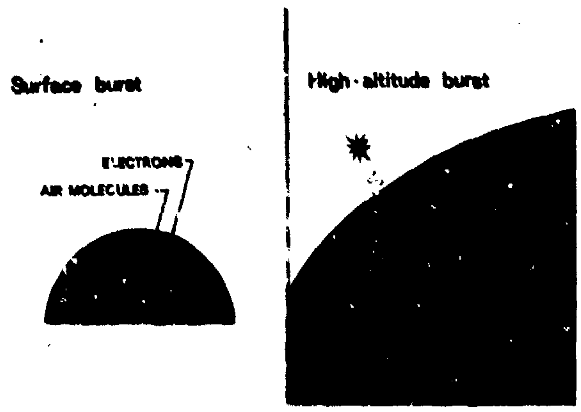
Discussion of close-in or source region effects is limited to a brief discussion of what System Generated EMP (SGEMP) and Internal EMP (IEMP) are. Discussions of how these environments couple to systems or their effects on systems are beyond the scope of this course. Further, other close-in effects of time-varying air conductivity, etc., are beyond the scope of this course. Consequently, the discussions in the remaining sections are restricted to the radiated EMP.



An energy flow diagram which illustrates the transformation of energy involved in the process of EMP generation is presented in the figure. The energy released from a nuclear detonation in the form of prompt gamma rays interacts with the earth's atmosphere to produce electrons and positive ions. X-rays can also produce an EMP at certain altitudes/burst regimes, but this phenomena is not considered here.



This separation of charge produces an electric field. The movement of electrons constitutes an electric current which has an associated magnetic field. These fields are coupled and radiate providing the proper conditions of asymmetry or geomagnetic field interaction exit. In the case of the near surface burst, the net charge separation caused by asymmetry of the deposition region due to the earth/air interface results in the radiated fields. In an exoatmospheric burst, the earth's geomagnetic field bends the scattered electron current moving away from the burst point. This bending produces an efficient conversion of the energy of the moving electrons into a radiated electromagnetic pulse.

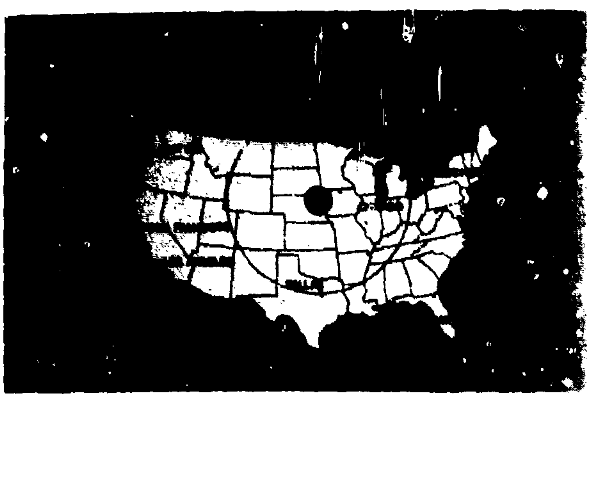


For both the high altitude and surface bursts, intense fields appear in what is called the deposition region. The deposition region for a surface burst is limited to about a two to ten kilometer diameter about the burst. For a high altitude burst, the source region can be on the order of 3000 kilometers in diameter. This deposition region extends from about 20 to 40 kilometers in altitude. In the deposition region, weapons effects other than EMP (such as radiation, blast, thermal, etc.), must also be considered.

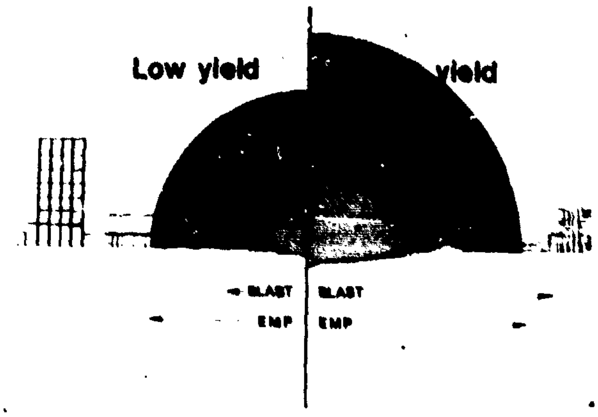
The electric field in the deposition region can be in excess of 100,000 volts/meter. Less intense fields appear outside the source region (the radiated fields). The maximum fields on the earth's surface due to an exoatmospheric burst are of the order of 50,000 volts/meter. For a near surface burst, the radiated fields outside the deposition region can be on the order of 1000 - 5000 volts/meter. The radiated fields from the air burst are less intense than the cited cases due to the near symmetry of the deposition region.

The radiated EMP has a fast rise time and thus has a spectrum which occupies a major portion of the communications band. Significant spectral energy is present in the exoatmospheric EMP over the frequency range of a few kilohertz to 100 megahertz.

In the case of an exoatmospheric burst, a significant overpressure pulse does not exist near the surface of the earth. Almost all of the other prompt weapon effects are diminished by the atmosphere, so that the most significant prompt weapon effect is the EMP. As noted previously, the source region can be quite large, in the order of 3000 km in diameter. As a consequence, the radiated fields from this source region can cover a substantial fraction of the earth's surface.



The size of the deposition region is confined by the atmosphere for the near surface burst. However, the pressure pulse which causes structural damage, is not similarly restricted but is instead proportional to the weapon yield. Thus, in the case of soft systems from an overpressure viewpoint, the most severe EMP exposure at otherwise survivable contours is associated with the low-yield near surface burst. If weapon systems have high overpressure survivability criteria, then close-in EMP for surface bursts must be considered.

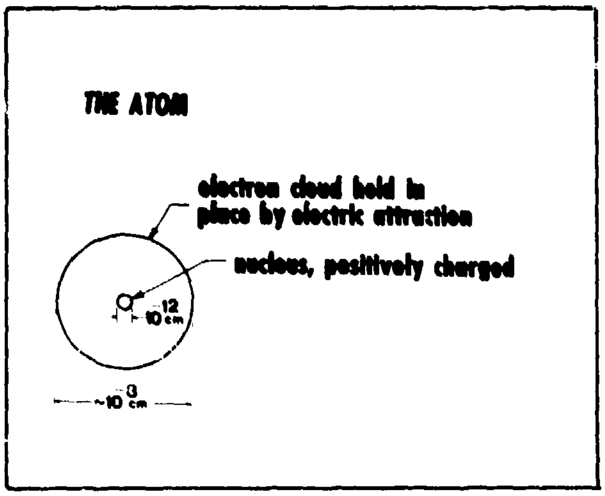


Another nuclear weapon effect often confused with EMP is communications "blackout". This effect also results from the ionization of the atmosphere due to the nuclear detonation. In the case of "blackout" this ionization affects the propagation path, that is, it changes the effective height of the ionosphere or increases the attenuation, such that effective communications are lost. "Blackout" does not interact with the terminal equipments resulting in damage or upset as is the case with EMP. "Blackout" is also a late-time (seconds to hours) effect; EMP is a prompt effect. "Blackout" will not be considered further in this discussion.

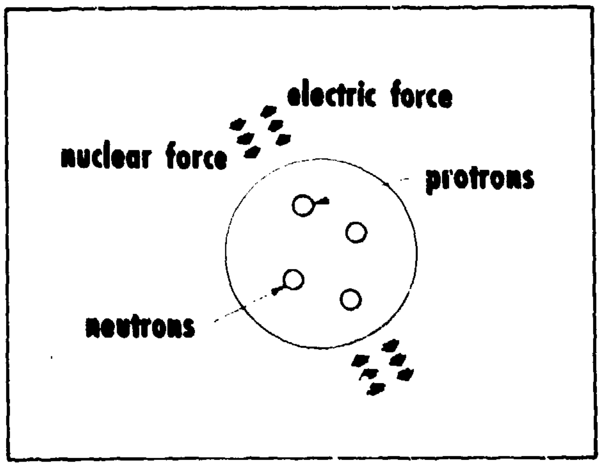
3.2 BASIC ATOMIC AND NUCLEAR PHYSICS

To understand the basic energy transformations from nuclear energy to EMP energy requires a review of a few basic facts from atomic and nuclear physics. The discussion which follows will consist of a review of atomic structure and energy levels within the atom, nuclear fission and the production of gamma rays, and the interaction of the gamma rays with the atmosphere and the conditions for radiation of electromagnetic fields.

The structure of an atom can be visualized in the familiar form of a small, positively charged nucleus surrounded by an electron cloud. The electron cloud is held in place by the electric coulomb attraction between the nucleus and electrons. The nucleus is on the order of 10^{-12} cm in diameter, while the electron cloud is about four orders of magnitude larger.

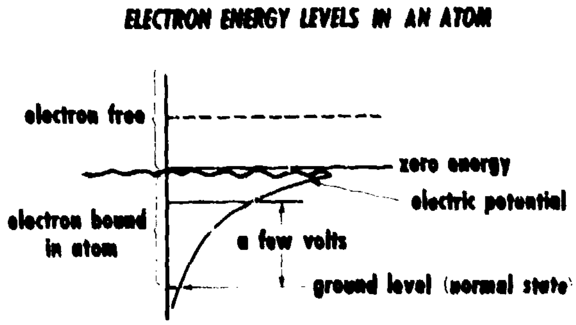


The nucleus is made up of protons (single positive charge) and neutrons (zero charge). The nucleus is held together by intense, short-range forces which are not yet completely understood. These nuclear forces are so strong that they overbalance the electric coulomb repulsion of the protons for each other.



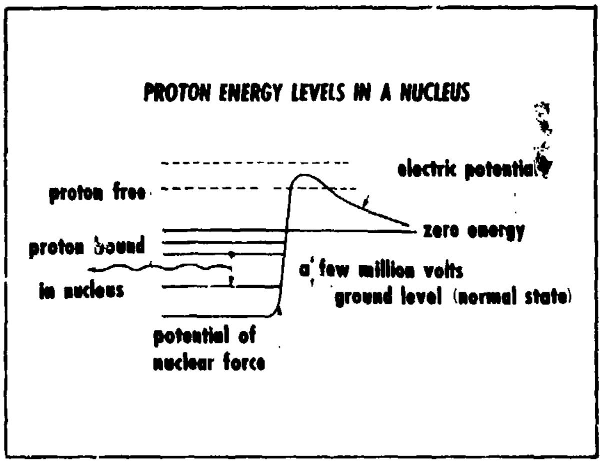
Atoms and nuclei exist in states having certain discrete energies. This is the basis of the quantum theory developed by Bohr, Planck, and many others. This theory gives a complete explanation of chemistry and atomic physics. There can be no doubt of its essential correctness.

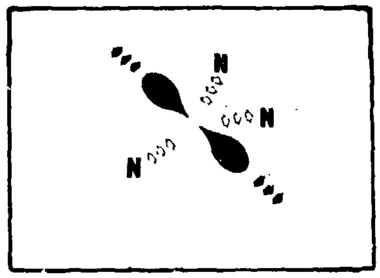
The energy difference of electron levels is of the order of a few electron volts. An electron can fall from one level to a lower one, at the same time emitting a quantum of light (photon). The energy lost by the electron is carried away by the light quantum. For example, "green" light quanta have energy of about 2.5 electron volts. An electron volt is the kinetic energy gained by an electron when it is accelerated through a potential of one volt.



The energy difference of proton levels in a nucleus is of the order of a few million electron volts. A proton can fall from one level to a lower one, at the same time emitting a gamma ray (photon). Again energy is conserved. There is no difference between gamma rays and light quanta except that gamma rays have about a million times more energy per quantum.

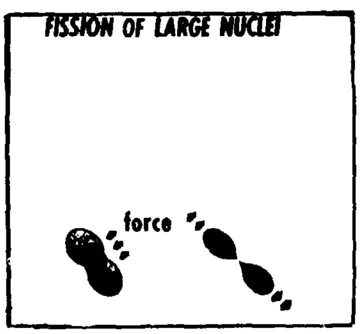
Gamma rays are the principal cause of radiated region EMP. About 0.1% of the energy of a typical nuclear bomb appears as prompt gamma rays. How does this happen? Let us consider the fission process in order to answer this question.





The protons in a nucleus repel each other electrically. In a spherical nucleus the electric repulsion is overbalanced by the nuclear forces. Ordinary nuclei are held spherical by a surface tension.

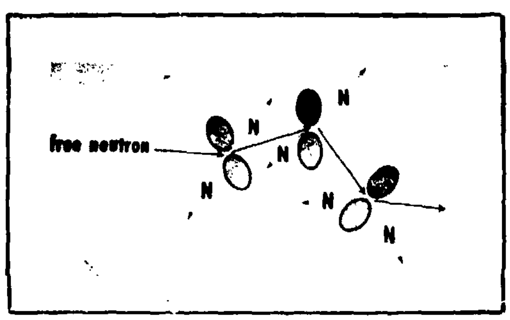
In large nuclei, the surface tension is not strong enough to keep the nucleus spherical. The electric repulsion tends to make the nucleus elongate, eventually dividing into two roughly equal parts. This is the fission process. This is one of the basic reasons why nuclei larger than uranium do not exist in nature; they are unstable against fission.



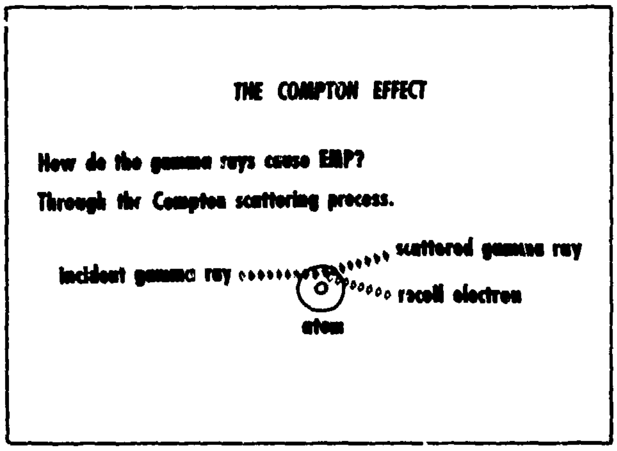
Some nuclei, like U^{235}, are just on the verge of being unstable. If such a nucleus is hit by a free neutron, it may undergo fission. The fission process is not neat and tidy as a few free neutrons get lost in the rush. These freed neutrons may hit other U^{235} nuclei, causing them to fission.

We will now consider the concept of a chain reaction. On the average, a freed neutron travels about 10 cm before striking another U^{235} nucleus and making it fission. If the piece of U^{235} is too small (sub-critical) the freed neutrons will escape without causing further fissions. If the piece of U^{235}, is large (super-critical), the number of fissions will grow exponentially with time, with the number of fissions proportional to $e^{t/\tau}$. The enfolding time τ is approximately equal to the travel time of a freed neutron before hitting another U^{235} nucleus. This time is in the order of 10 nanoseconds. The EMP can have a comparable rise time.

Where do gamma rays come from? The fission fragments are usually not born in their ground levels. Free neutrons collide with other nuclei in the bomb or air or earth, knocking some of these into levels above the ground level. Gamma rays are then emitted in transitions back to the ground level.



How do the gamma rays cause EMP? The answer is through the Compton interaction process. Compton discovered that photons can collide with electrons, knocking them out of the atoms in which they were originally bound. These Compton collisions are somewhat like the collision of a moving billiard ball with one at rest. The recoil electron, like the ball originally at rest, goes predominantly forward after the collision. Thus, a directed flux of gamma rays produces, by Compton collisions, a directed flux of electrons. This constitutes an electric current, which generates the EMP.



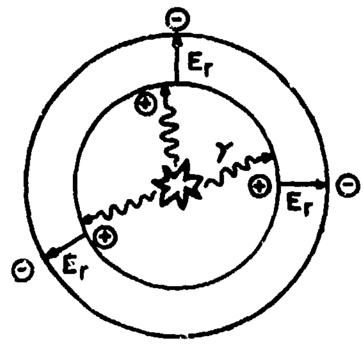
We can now consider some order of magnitude estimates of the energy involved at each state of the EMP generation process. A one megaton bomb releases 4×10^{15} joules of energy. About 0.1% of this energy, or 4×10^{12} joules, may appear as prompt gamma rays. This amount of energy is equivalent to that produced by a hundred megawatt power plant running for about 11 hours. A fair fraction, about one-half of this, goes into the Compton recoil current. Fortunately, most of this energy goes into heating air rather than into the EMP. About 10^{-3} of the gamma energy goes into EMP; thus giving about 10^{-6} of the bomb energy going into the EMP.

3.3 EMP GENERATION

The previous section showed how a nuclear detonation produces the gamma rays and how these in turn produce the Compton electrons. These basic physics principles hold for all burst locations. To generate a radiated EMP, however, other mechanisms exist which differ with burst location. This section will discuss generation of a radiated EMP and its characteristics for the three burst locations.

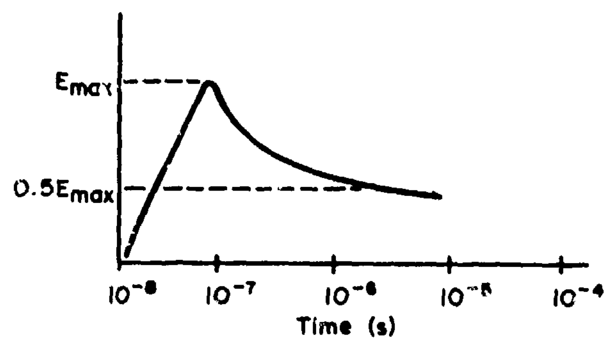
Deposition Region Fields

The gamma rays emitted by the nuclear detonation are nearly symmetrical, any anisotropy of the emission of the gammas is small and of short duration compared with other factors. The result is the Compton electrons created move radially away from the burst point and there exists a radial symmetric current distribution. The scattering of the electrons leaves behind the heavier positively charged parent molecule. This separation of charge produces a strong radial electric field as shown in the figure.



CHARGE SEPARATION MODEL

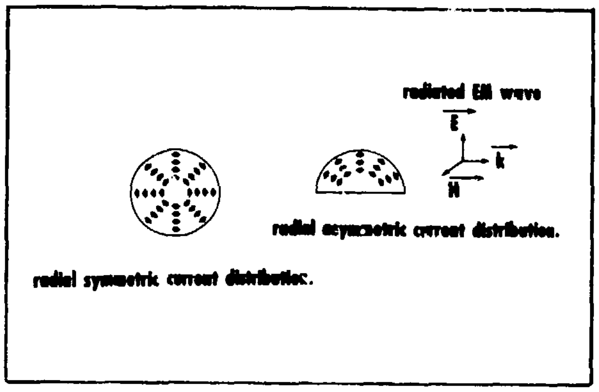
For this symmetrical condition, no magnetic field is generated and therefore the deposition region fields are nonradiating fields. In the deposition region in early time, the Compton electron current dominates and the radial electric field rises rapidly. Due to additional collisions, the Compton electrons produce secondary electrons which further ionize the air and increase its conductivity. These secondary electrons under the influence of the radial electric field move and produce a conduction current which tends to reduce or limit the local field resulting in the plateau at later time.

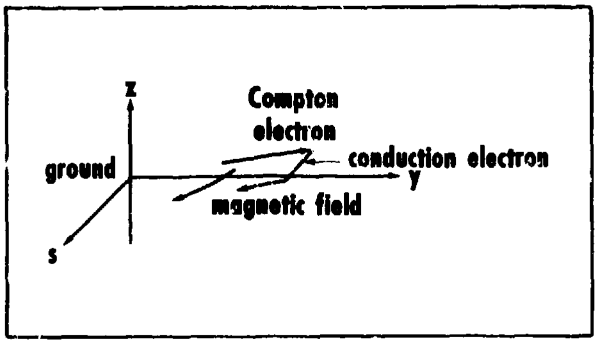


GENERAL TIME WAVEFORM OF E_r IN DEPOSITION REGION

Near Surface Burst EMP

A completely symmetrical system of radial currents produces neither magnetic nor radiated fields. The departure from symmetry, in the case of the near surface burst (less than 2 kilometers above the earth), is provided by the earth/air interface, resulting in a nearly hemispherical deposition region.



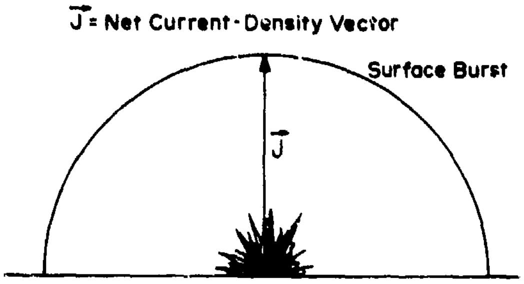


The deposition region is limited in size (3 to 6 km) by the air density and hence the distance the gamma rays can travel. The average distance a gamma ray travels before making a Compton collision is about 200 meters at sea level, although a few may go as far as a few kilometers. The Compton electrons travel outward only a few meters before being stopped by the air. The deposition region is thus more dependent on the absorption of gammas in the atmosphere than on the yield of the weapon.

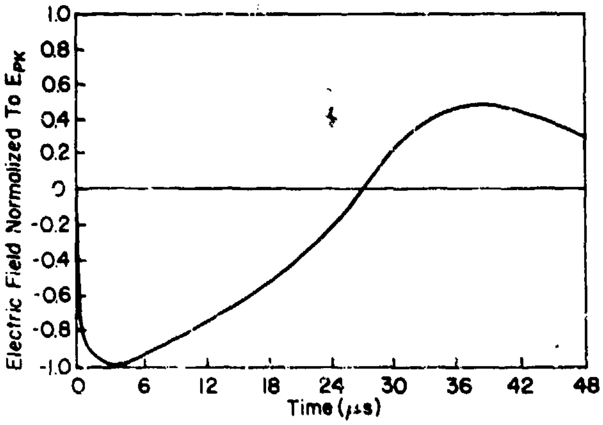
Within the deposition region, as discussed previously, a strong radial field exists. The radial field has a peak amplitude of approximately 100 kV with a rise time of 10 nanoseconds. The earth tends to short out the radial electric field near it since it is normally a better conductor than air. Near the ground the conduction electrons find an easier path back to the positively charged center by flowing down to the ground and back towards the burst point. The result is a current loop which generates an azimuthal magnetic field.

A vertical electric field is required in connection with the vertical component of conduction current. This vertical electric field can be regarded as connecting Compton electrons in the air with their image charges in the ground. Thus, for a near surface burst, the principle fields in the deposition region are a radial and vertical electric field, and an azimuthal magnetic field.

When viewed from a large distance from the burst point, the net fields are the vertical electric and azimuthal magnetic fields. A vertical dipole can be used as a source model for the radiated EMP in this case. The dipole model is depicted in the figure.



ELECTRIC-DIPOLE MODELS OF SURFACE AND AIR-BURST RADIATED EMP (ADAPTED WITH THE PERMISSION OF THE DEFENSE NUCLEAR AGENCY)



GENERALIZED TIME WAVEFORM OF SURFACE-BURST RADIATED ELECTRIC FIELD (ADAPTED WITH THE PERMISSION OF THE DEFENSE NUCLEAR AGENCY)

The radiated fields are characterized by an oscillating signal. The magnitude of the radiated field falls off inversely proportional to distance from the burst. A rough estimate of the peak electric field strength (E_{pk}), where peak is the largest negative value, is given by:

$$E_{pk} = \frac{10^7}{R} \text{ volts/meter}$$

where R = radial distance from the burst point.

A generalized time waveform is shown. The Fourier transform of this waveform shows the bulk of the energy lies below 1 megahertz. The peak amplitude at a distance of 10 km is about 1000 volts/meter.

The radiated field from a near surface burst is predominantly vertically polarized at the earth's surface.

The relative magnitude of the deposition region fields to the radiated fields, indicate the near surface burst EMP is a serious threat to most systems within the deposition region. Outside the deposition region, the EMP is a principal threat to systems which respond to very low frequencies or have very large energy collectors.

Air Burst EMP

An air burst is defined as a burst occurring at an altitude of 2 to 20 kilometers above the earth. The EMP from an air burst is characterized by a strong radial electric field in the deposition region and a weak radiated field.

The air density, in the range of 2 to 20 kilometers, is relatively dense and quite uniform. This relatively dense air results in a relatively short range for the gamma rays and Compton electrons as in the case of the near surface burst. Consequently, no appreciable turning of the electrons takes place in the geomagnetic field (see the discussion on exoatmospheric EMP), and therefore, this mechanism does not result in a radiated EMP.

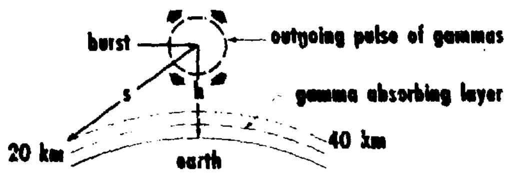
Since the air density is also quite uniform, only a small net current asymmetry exists resulting in only a weak radiated EMP. A dipole model, as used for the near surface burst, can also be used here. The radiated fields from the air burst are similar in shape to those of the near surface burst, with the peak fields at least one order of magnitude less than those of the near surface burst.

Due to the weak radiated fields, the EMP from an air burst is a principle threat to systems which may be within the deposition region, in which case, the other nuclear weapons effects must also be considered.

Exoatmospheric Burst EMP

An exoatmospheric burst is defined as one which takes place at an altitude greater than 40 kilometers above the earth. A typical exoatmospheric burst geometry is depicted in the figure.

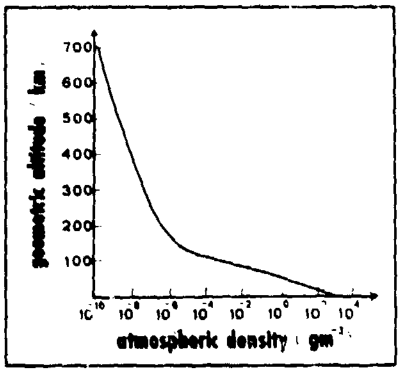
EMP FROM HIGH ALTITUDE BURSTS
EXAMPLE OF GEOMETRY BEING CONSIDERED



h = height of burst ≈ 400 km
s = distance to horizon ≈ 2,250 km

A high altitude burst illuminates large geographical regions with gamma rays

Above about 40 kilometers altitude, the atmospheric density is sufficiently small that the high energy gammas are not affected appreciably. The atmospheric density is large enough that the gammas are absorbed by Compton scattering below 40 kilometers. The gamma absorption is nearly complete by the time they reach 20 kilometers altitude. The deposition region for a high altitude burst is thus between about 20 to 40 kilometers, which is approximately 65,000 to 130,000 feet.



The Compton electrons are scattered forward (downward in this case), within the deposition region as was the case for both the near surface and air bursts. These Compton electrons produce a system of radial currents. As stated previously, a symmetrical system of radial currents will not produce a radiated field. In an exoatmospheric burst situation, there are two mechanisms which produce the radiated field: (1) the asymmetry due to the space/atmosphere interface, and (2) Compton electron turning due to the earth's geomagnetic field. It is the Compton electron turning which dominates since the electrons can travel greater distances, and thus have more time to interact with the geomagnetic field due to the reduced atmospheric density.

The outgoing gammas from the burst form a spherical shell which expands with the velocity of light. Since most of the gammas are emitted in about 10 nanoseconds, the thickness of the shell at any instant is a few meters. When the gamma shell begins to intersect the absorbing layer of the atmosphere, the Compton scattering process begins.

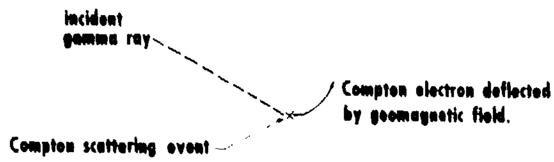
At the altitude of the deposition region, the stopping range of Compton recoil electrons is of the order of 100 meters. In traveling this distance, the Compton electrons are strongly deflected by the geomagnetic field with a gyro radius of about 100 meters. The Compton recoil current therefore has strong components in directions transverse to the gamma propagation direction. This transverse current radiates an electromagnetic wave that propagates in the forward direction.

The outgoing wave keeps up with the gamma shell and is continually augmented by the transverse Compton current until the gammas are all absorbed. Then the electromagnetic wave goes on alone as a free wave or pulse. Secondary electrons produced by the Comptons make the air conducting. This conductivity attenuates the electromagnetic pulse. The amplitude of EMP is determined by a balance between:

1. Increase due to transverse Compton current.

2. Attenuation due to conductivity.

MECHANISM OF EMP GENERATION

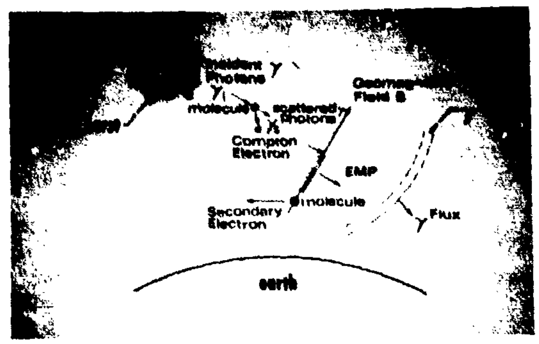


A summary of the process of EMP generation from a high altitude burst can now be given along with important details on the directional dependence.

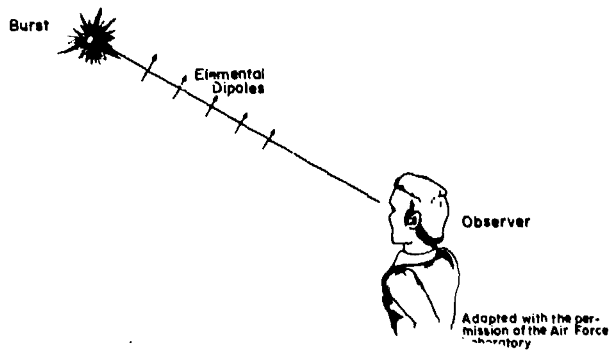
Gamma rays are scattered from molecules with the emission of Compton electrons in the forward direction with energies on the order of 1/2 million electron volts. The motion of the Compton electrons is modified by the geomagnetic field. They follow a spiral path about the magnetic field lines until they are stopped by collision with atmospheric molecules. As the Compton electrons collide with atmospheric molecules, further ionization occurs and the conductivity increases. The propagation of the EMP depends on the conductivity of the region through which it passes. Dispersion, attenuation, and reflection may occur. The circular component of the Compton electrons represents a magnetic polarization, and the linear component represents an electric polarization.

Both the magnetic and electric polarizations vary with time and, consequently, can radiate electromagnetic energy. This can also be viewed as a collective flow of electrons along the field which radiates in the transversal direction.

Compton electrons that move parallel to the geomagnetic field are not deflected. Thus, the EMP amplitude is small in two directions along the geomagnetic field line passing through the burst point. The EMP amplitude is a maximum on those rays from the burst point which run perpendicular to the geomagnetic field in the deposition region.



The transverse current components and the resulting radiated fields can be understood by considering a system of elemental dipoles along the line-of-sight from the observation point to the burst point. Each dipole radiates an electromagnetic field (the normal "donut" radiation pattern of a dipole) which add in phase since the EM fields, gammas, and Compton electrons all travel at or near the speed of light.



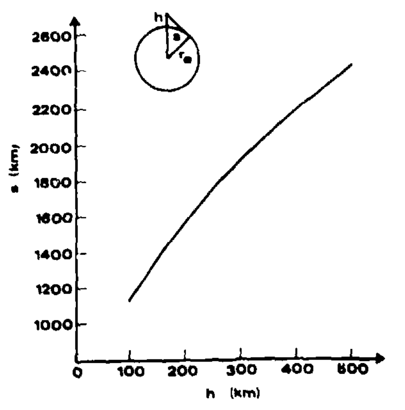
PHASED MAGNETIC DIPOLE ARRAY MODEL

The radiated fields are high intensity fields, have short rise times (wide spectrum) and cover a large area of the earth's surface because of the height and large extent of the deposition region. The geographical coverage as a function of the height of burst can be obtained by considering the tangent radius.

The tangent radius is the arc length between the line from the earth's center to the burst point and the line from the earth's center to the point where a line from the burst point is tangent to the surface of the earth. For a burst at 300 kilometers, the tangent radius is 1920 kilometers, or about 1200 miles. The EMP from the burst can cover this region.

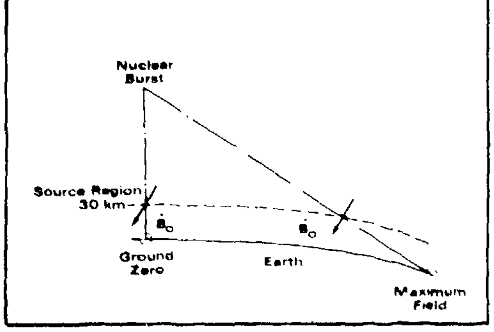
It should be noted that the fields do not end at the tangent radius. For surface systems, this is the farthest point of interest, in most cases, but low amplitude fields may be propagated beyond this point by ducting. For airborne systems, however, the coverage may be much farther depending on the height of the point of observation. As long as there is line-of-sight to the burst point, the EMP will be present.

TANGENT RADIUS FOR A HIGH ALTITUDE BURST

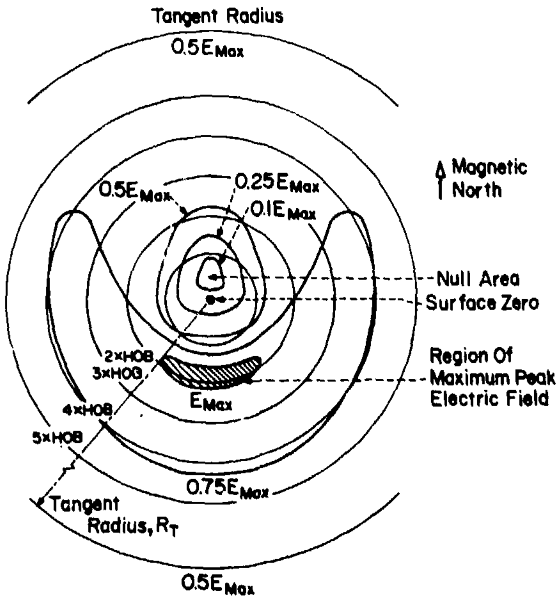


The EMP time waveform varies considerably over the area of coverage. Three cases will be considered: (1) near surface zero, (2) the maximum field point (where a line from the burst point to the maximum field point is orthogonal to the geomagnetic field lines in the deposition region), and (3) at the tangent radius point.

HORIZONTAL VARIATION OF EMP FROM A HIGH ALTITUDE BURST



At surface zero, the fields have reduced amplitude (0.25 E_{max}), short rise time (approximately 5 nanoseconds), and time to half value of about 20 nanoseconds. At the maximum field location, the rise time is about 10 nanoseconds and time to half value is about 50 nanoseconds. At the tangent radius, the fields are about 0.5 E_{max}, and rise time to half value greater than 10 nanoseconds and 200 nanoseconds respectively. This distribution, as a function of height of burst and geographic position on the earth, can be determined from the figure. The peak field at any point can be found by aligning magnetic north on the figure with magnetic north at surface zero.



VARIATIONS IN HIGH ALTITUDE EMP PEAK ELECTRIC FIELD ON SURFACE OF CONTINETAL UNITED STATES (ADAPTED WITH THE PERMISSION OF THE DEFENSE NUCLEAR AGENCY)

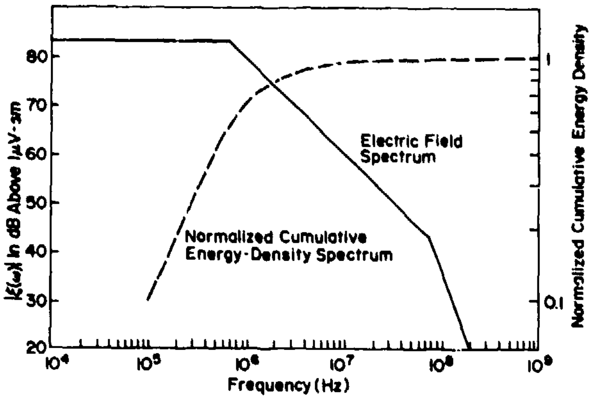
As a result of this wide variation on the earth's surface and the inability to predict the burst location, the time waveform used preserves the important characteristics of the three waveforms discussed. The composite waveform, therefore, maintains the fast rise time (near surface zero rise time), slow decay (near tangent radius decay), and maximum peak amplitude. The spectrum contains all frequencies of interest. The composite waveform, shown in the figure, can be approximated analytically by a double exponential.

$$E(t) = 5.25 \times 10^4 \left[e^{-\alpha t} - e^{-\beta t}\right]$$

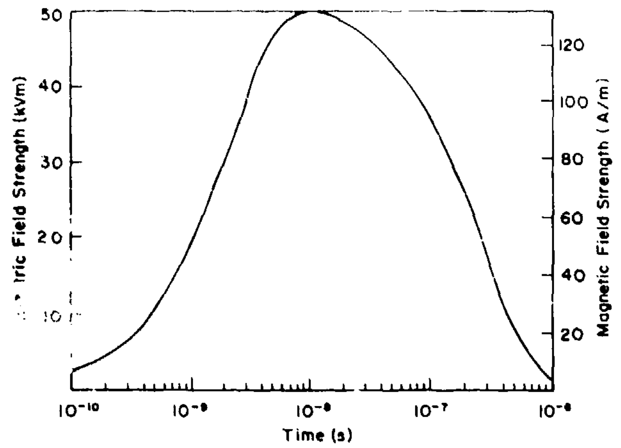
where

$$\alpha = 4 \times 10^6$$

$$\beta = 4.76 \times 10^8$$



HIGH-ALTITUDE EMP SPECTRUM AND NORMALIZED ENERGY DENSITY SPECTRUM



GENERALIZED HIGH-ALTITUDE EMP ELECTRIC-AND-MAGNETIC-FIELD TIME WAVEFORM

The spectrum of this pulse can be obtained by taking the Fourier transform of the time waveform. The significant frequencies extend out to about 150 megahertz with the bulk of the energy (99.9%) below about 100 megahertz.

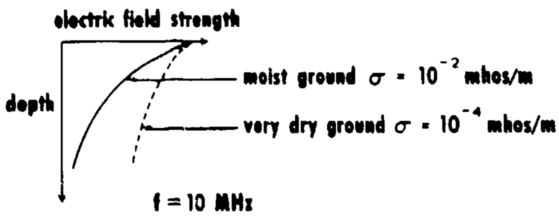
The polarization of the EMP depends on the locations of both the burst and the observer, and the orientation of the geomagnetic field. The direction of the electric field is normal to both the geomagnetic field at the observer's location, and the direction of incidence. The direction of incidence is radially outward from the burst. For the Continental U.S., a typical dip angle for the geomagnetic field is about 67 degrees.

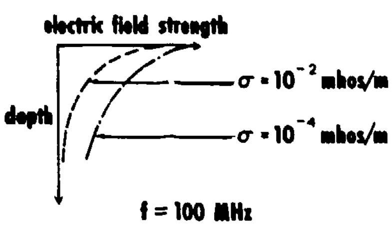
For this dip angle, and assuming the geomagnetic field lines run north and south, the EMP polarization is horizontal for bursts north or south of the observer. For bursts east or west of the observer, the polarization departs from horizontal by 23 degrees or less. Therefore, the principle polarization for systems based in the U.S. is horizontal.

3.4 EARTH EFFECTS ON TOTAL FIELDS

So far we have discussed the free field which impinges on the surface of the earth due to the radiated EMP. Since these are electromagnetic waves, they will penetrate the surface of the earth and be reflected by it. The amount of penetration or reflection is a function of the frequency (spectrum of the pulse), the polarization, direction of incidence, and earth parameters.

The reflection and transmission coefficient equations are presented in Section V. It should be noted that a phase reversal takes place for the component of the electric field parallel to the conducting surface but not for the normal to the surface component. Near the surface, therefore, for the parallel component, the electric field is reduced, but the magnetic field is increased. Electromagnetic fields also diffuse into the ground. The penetration depth depends on ground conductivity and frequency component considered. A greater ground conductivity leads to a smaller penetration depth. The high-frequency components penetrate less deeply than the low-frequency components of the pulse.

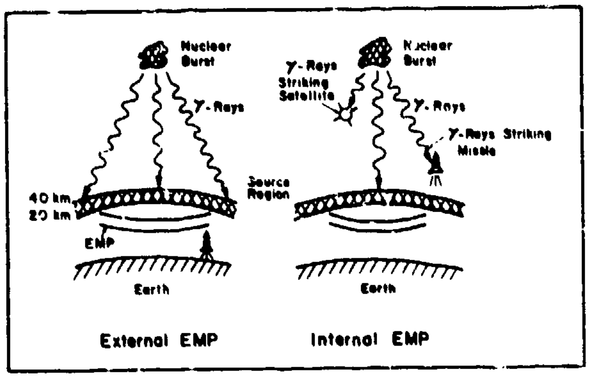




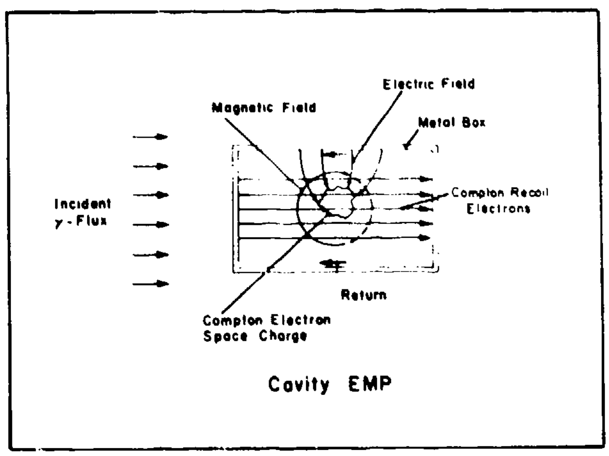
3.5 SYSTEM GENERATED AND INTERNAL EMP

While not the subject of this course, a few words are in order in the area of system generated EMP (SGEMP) and internal EMP (IEMP).

Internal EMP is EMP induced directly in a system by gamma radiation (or X-rays) striking the system. In the case of internal EMP, Compton recoil electrons are produced by interaction of incident gamma radiation with material in the system. This interaction is similar to production of Compton recoil electrons in the atmosphere in the case of external EMP. However, the way in which Compton recoil electrons produce internal EMP differs from that of the Compton recoil electrons in external EMP.

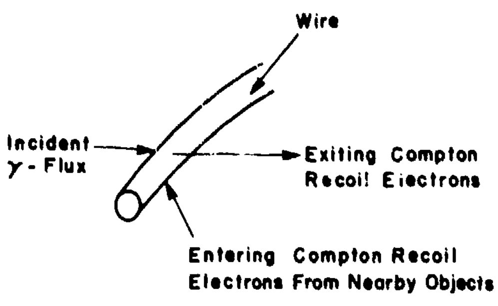


Internal EMP can arise in different ways. One type may be called "cavity EMP" and arises from Compton recoil electrons being driven across the volume of an enclosed cavity. Both electric and magnetic fields are generated. These fields may ring at the cavity frequencies. The fields will induce voltages in circuits contained in the cavity. If the cavity contains air, it may become electrically conducting. This affects the time behavior of the fields.



Another internal effect (SGEMP) is "Compton charging." Any object in a gamma flux has some Compton electrons knocked out of it and receives Compton electrons knocked out of other nearby objects. These will usually not balance. The charge added at various locations in a conducting circuit will flow away through the circuit, thus generating an extraneous signal. The charge acquired by insulation may leak away very slowly. In general, the thicker the object, the more charge it acquires.

We will not discuss further cavity EMP or Compton charging in this course. This does not mean that they are unimportant, but rather that they are another problem area sufficiently different from EMP to be considered in a separate course.



Compton Charging

This very brief discussion does point out that IEMP and SGEMP are of primary concern to space systems or to systems which are within the deposition region of the burst. These effects both require direct exposure to the incident gamma flux which occurs only in the deposition region for near surface or air bursts. Since the gamma rays are not absorbed in space, space systems even at considerable distance may be exposed to the gamma flux.

SECTION IV

EMP INTERACTION AND COUPLING ANALYSIS

4.1 INTRODUCTION

To be able to assess the effects of EMP on a system, the way EMP interacts with the system and the extent of this interaction must be understood. The discussion on interaction will consider, in a qualitative manner, the important aspects of electromagnetic coupling. The discussion of coupling analysis will review methods for obtaining quantitative estimates of this coupling.

In discussing interaction, we will consider the principles by which energy is transferred from the electromagnetic field to collectors. The effects of the earth and other structures which may alter the incident electromagnetic field will be considered. The mechanisms by which shields and cables work and by which electromagnetic energy may penetrate into the system interior will also be discussed.

The discussion on coupling analysis will consider the basic methodology for predicting, in a quantitative way, the energy transferred by an electromagnetic field to a collector. Examples of the application of this methodology to such collectors as antennas, cables, and conducting structures will be discussed.

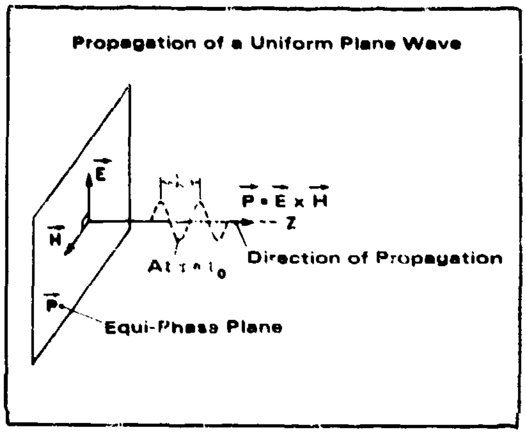
4.2 EMP INTERACTION WITH SYSTEMS

The interaction between an electromagnetic wave, such as EMP, is highly dependent on the characteristics of the wave, the characteristics of the system, the characteristics of the propagating media, and the media or structures in the vicinity of the system of interest. The role each of these plays will be discussed in this section.

The Plane Wave

Outside of the deposition region for a near surface nuclear burst, the propagating EM wave can be modeled as a plane wave. For the exoatmospheric burst, on the surface of the earth, the EMP can also be locally modeled as a plane wave. Modeling the EMP radiated wave as a plane wave, or a uniform plane wave, implies very well defined characteristics.

For a sinusoidally time varying field to be plane, the electric and magnetic fields must be mutually orthogonal and in phase everywhere in a plane perpendicular to the direction of propagation. In addition, if the magnitude of the electric and magnetic fields are independent of position in the plane (i.e., the magnitude of the electric field and magnetic field at any instant of time is the same anywhere in the plane), the wave is defined as a uniform plane wave.



Propagation of a Uniform Plane Wave

The wavelength, λ, of a monochromatic (single frequency sinusoidal) plane wave is the separation distance between planes in which the E vector in one plane is in time phase with the E vector in the reference plane.

The uniform plane wave has other properties that make it convenient to work with. One of these is if the wave is propagating through a nondispersive (lossless) medium, the electric and magnetic fields are related through the spatial impedance of the medium, called the intrinsic impedance of the medium (η).

$$\frac{|\vec{E}|}{|\vec{H}|} = \eta.$$

$$\eta = \sqrt{\mu/\varepsilon} = 377\ \Omega \text{ in air or free space}$$

The velocity of propagation of a plane wave through a nondispersive medium is the velocity of light through that medium. If C is the velocity of light, then

$$C = \frac{1}{\sqrt{\mu\varepsilon}} = 3 \times 10^8 \text{ meters/sec in air of free space}$$

If the plane wave is monochromatic, the phase shift per unit length due to the finite propagation velocity is given by the phase constant, β.

$$\beta = \frac{2\pi}{\lambda} = \frac{\omega}{C} = \omega\sqrt{\mu\varepsilon} \text{ radians/meter}$$

$$\omega = 2\pi f = \text{radian frequency.}$$

The average power density of a uniform plane wave (single frequency CW) is given by the magnitude of the Poynting vector $\vec{P}$.

$$|\vec{P}| = |\vec{E} \times \vec{H}| = |\vec{E}|\ |\vec{H}| \sin\theta \text{ watts/m}^2$$

$$\theta = \text{angle between } \vec{E} \text{ and } \vec{H}$$

If $\vec{E}_m$ and $\vec{H}_m$ are the maximum values of the sinusoidal fields of a uniform plane wave, then the average value of the power density is given by

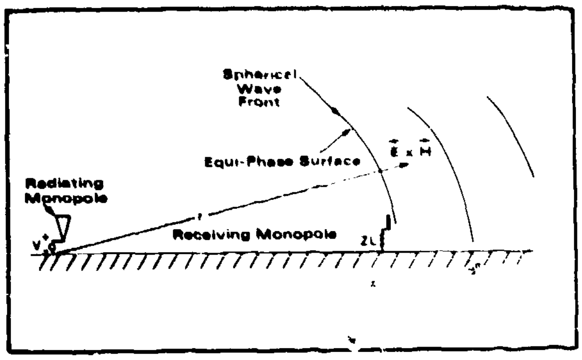
$$|\vec{P}|\text{ avg} = \left|\frac{\vec{E}_m}{\sqrt{2}} \times \frac{\vec{H}_m}{\sqrt{2}}\right| = \frac{|\vec{E}_m|\ |\vec{H}_m|}{2}$$

$$= \frac{|\vec{E}_m|^2}{2\eta} \text{ watts/m}^2$$

When a uniform plane wave is a pulsed field in the time domain, such as the EMP from a nuclear detonation, it may be considered as consisting of many frequency components constituting the frequency spectrum of the fields. In this case, the $\vec{E}$ and $\vec{H}$ fields are functions of the radian frequency (ω). The power density of a uniform plane wave for the pulsed field is defined as a spectral power density and is given by:

$$P = \frac{|E(\omega)|^2}{2\eta} \text{ watts/m}^2\text{Hz}$$

A perfectly uniform plane wave cannot exist in nature; however, it is an extremely useful and reasonably accurate approximation for such wave fronts as spherical and cylindrical waves which are common in nature, providing the proper conditions are met. For example, the free-space, far-zone field of a radiating monopole antenna is a spherical wave.



To a distant receiving monopole antenna, the wave appears as if it were a uniform plane wave since the receiving antenna is sampling only a small portion of the spherical wave front. This same criterion must hold for the case of the EMP wave front; however, it is difficult to define the far fields from an EMP source region in terms of equivalent antenna dimensions of the source. In the case of a near-surface burst, the dipole model of the source region can be considered as a point source, so fields outside the source region can be considered planar even for systems a few kilometers long. In the exoatmospheric case, the source region is widely distributed. The fields on the surface of the earth are the result of an elemental dipole array model, the time of arrival difference, even for systems many kilometers in length, is so small that the wave can be approximated by a plane wave.

Fields Due to EMP

The preceeding paragraphs discussed the wave nature of the EMP and the relationship between the electric and magnetic fields. To understand the interaction of the EMP with systems, the time and frequency distribution of the pulse must be reviewed.

Consider the electric field time history for a test waveform which approximates a typical high altitude EMP in free space. The waveform exhibits an extremely fast rise time (~ 5 ns to 10 ns between the 0.1 and 0.9 E_{max} amplitudes), and a slow decay (~ 500 ns to 1 μs, 0.9 to 0.1 E_{max}). For analytical purposes, the waveform usually can be approximated by a superposition of two exponential functions as

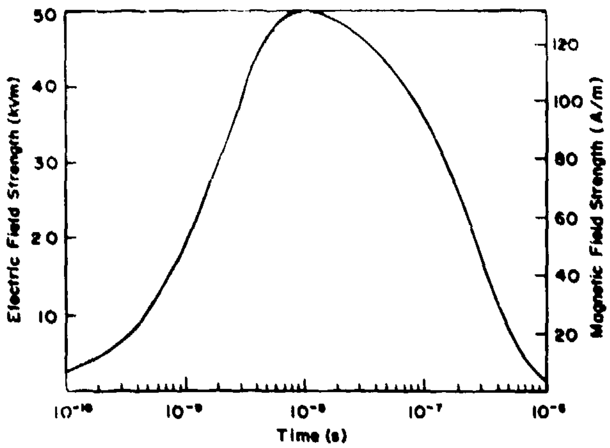
$$E(t) = A\,(e^{-\alpha t} - e^{-\beta t}) \text{ volts/m}$$

Assuming the wave to be a plane wave in free space, the magnetic field has the same time dependence (shape) and is related to the electric field by

$$H(t) = \frac{E(t)}{\eta_o} \text{ amp/m}$$

where

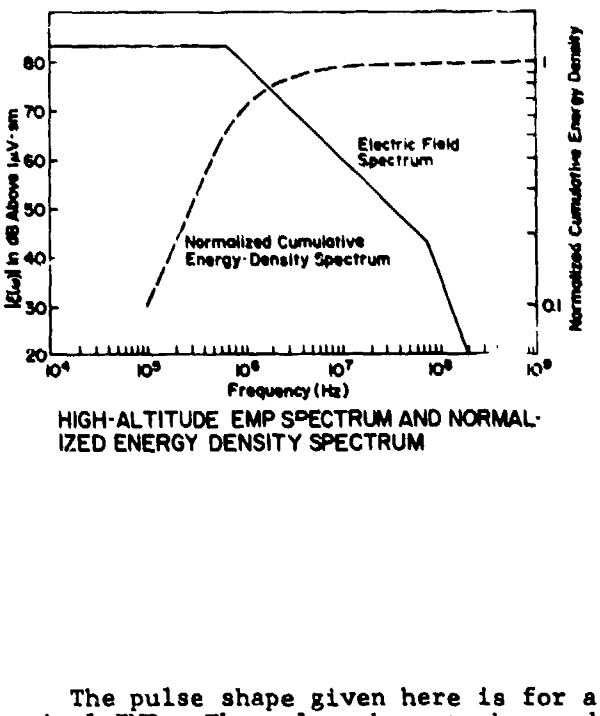
η_o = 377 ohms (intrinsic impedance of free space)



GENERALIZED HIGH-ALTITUDE EMP ELECTRIC-AND-MAGNETIC-FIELD TIME WAVEFORM

This waveform, as presented, has a dc component which cannot exist in a radiating wave. Unless this is recognized, some non-physical results can be obtained in coupling analyses. The portion of the spectrum below the audio range (10 kHz) is presently being investigated.

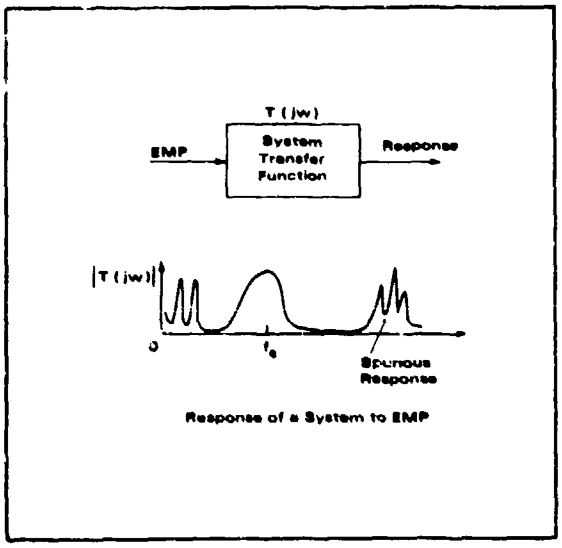
Many EMP collectors (antennas, aerial and buried cables, structures, etc.) are particularly frequency selective. It is, therefore, desirable to determine the EMP frequency spectrum by taking the Fourier transform of time waveform. The plot of the magnitude of a nominal EMP spectrum indicates two break points associated with the fall and rise time of the time waveform. It is important to note that the spectrum is almost constant below the HF range. Between the first and second break-frequency, the spectrum falls off at 20 dB/decade. Above the second break-frequency, it falls off at 40 dB/decade. Also shown is the cumulative energy density which is normalized to 0.9 joules per square meter. Approximately all the EMP energy (> 99 percent) is concentrated below 100 MHz.



HIGH-ALTITUDE EMP SPECTRUM AND NORMALIZED ENERGY DENSITY SPECTRUM

The pulse shape given here is for a nominal EMP. The pulse shape to be used for a particular scenario for a system analysis should be obtained from the Defence Nuclear Agency or the service lead laboratories.

The two factors of the EMP that contribute to the threat to electrical and electronic equipment are: (1) the large amplitude of the fields, and (2) the broad spectrum. EMP energy collection of a given system is a function of the system response to all frequencies in the EMP spectrum. The system transfer function may exhibit responses to both in-band and out-of-band frequency "windows". Out-of-band energy collection is a result of the system having a frequency response greater than needed to perform the system's function. This added frequency response can be deliberate (the designer wanting to have a better response than actually required), or can be non-deliberate such as spurious responses in a receiver. The main concern in the case of EMP is due to the broad spectrum; it is imperative in assessing the system vulnerability that the overall system response be determined and utilized in any analysis effort.

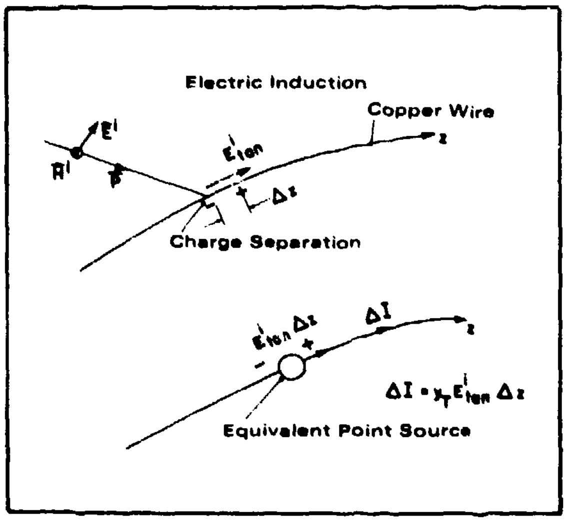


Response of a System to EMP

EM Induction Principles

There are three basic mechanisms by which the EMP energy couples to a conducting structure. These are: (1) electric induction which is the principle mechanism for linear conductors, (2) magnetic induction which is the principle mechanism when the conducting structure forms a closed loop, and (3) through the earth transfer impedance which is the principle mechanism for buried conductors.

A structure can collect energy from an impinging EMP field by electric induction. In effect, charges on a conducting surface are separated by the tangential component of the impinging electric field which results in current flow. The overall result is that a voltage source distribution is induced on the conductor consisting of incremental (point) voltage generators. The contribution of each point generator to the current at some point on the conductor is determined by its transfer admittance, which is a function of the source and observation points.



A coupling structure can also collect energy from an impinging EMP field by magnetic induction. The induced signal from an EMP is equal to the negative time rate of change of the incident magnetic flux. This is an empirically derived fundamental law known as Faraday's Induction Principle. The voltage induced in a loop by a uniform magnetic field is given by

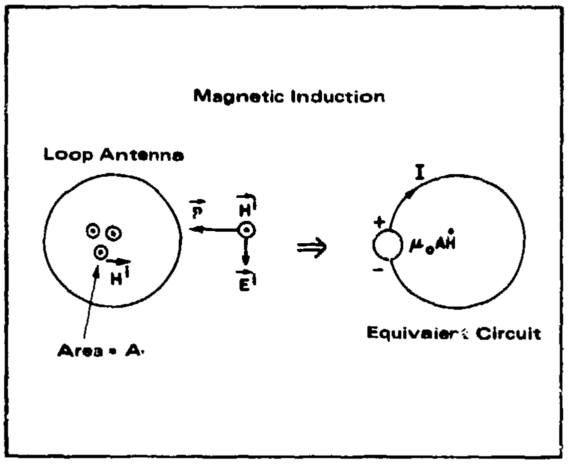
$$v(t) = \mu_o A \frac{d}{dt} H(t)$$

where

μ_o = $4\pi \times 10^{-7}$ is the permeability of free space

A = the area of the loop

H = The component of the magnetic field normal to the plane of the loop.



Another important coupling mechanism is that due to the I-Z drop which occurs whenever a coupling structure is imbedded in a lossy medium, e.g., earth, in which EM fields exist. Consider, for example, a bare wire conductor located a distance, d, below the surface of the earth and in the presence of an impinging EMP field as shown. The incident EMP field illuminating the surface of the earth causes conduction currents per meter width to flow in the earth. As a result of the current flow in the earth, a distributed I-Z drop appears along the wire which is equivalent to a voltage source distribution, causing current to flow in the wire conductor. An increment of voltage drop (which can be viewed as a point generator), ΔV, over a distance, Δz, along the conductor is given by

$$\Delta V = Z_T J \Delta z$$

where

J = total earth conduction current per meter width

Z_T = surface transfer impedance of the earth.

The surface transfer impedance is given by

$$Z_T \approx \frac{\alpha}{\delta} e^{-\alpha z} \exp\left[- d\sqrt{\alpha^2+\beta^2}\right]$$

where

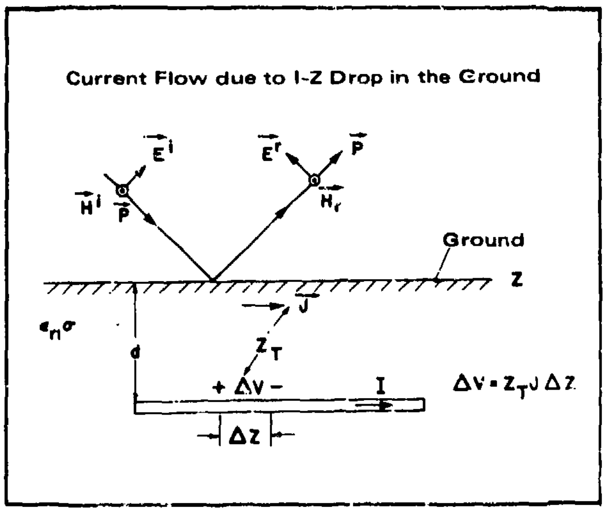
α = propagation constant of the earth

β = free-space propagation constant

δ = earth conductivity

d = depth of burial.

Transmission line theory may now be used to determine the actual current distribution induced on the wire.



Antennas

Electromagnetic coupling to antennas results from electric and magnetic induction. The induction principle which dominates for a given antenna depends on the antenna's geometrical configuration. At this time we will consider two basic, simple types: (1) the linear antenna (monopoles or dipoles), and (2) the loop antenna. The way energy couples to these simple antennas can serve as a basis for estimating the energy collection by complex antennas and the other conducting structures.

Linear Antennas

When an electromagnetic wave impinges on a linear antenna, the tangential component of the electric field induces charges on the antenna which tend to cancel the incident field. Since these charges are free to move, an antenna current results. In a time varying field, therefore, these charges are constantly in motion and an alternating current flows on the antenna and in the load.

Electric induction on a linear antenna can be viewed as an array of point voltage sources. To obtain the total voltage induced on the antenna, all of the point sources must be summed (integrated) over the length of the antenna. If the antenna is short (antenna length $\leq$ 1/6 the wavelength of the incident signal), it is not necessary to perform the integration. In the case of EMP, since it is a transient signal containing a wide frequency spectrum, the antenna length must be $\leq$ 1/6 the wavelength of the highest frequency component of interest in the spectrum. For typical EMP waveforms, the actual physical length must be less than approximately two (2) feet.

Considerable simplification results when the EMP coupling structure is electrically small, since, for such structures, simple equivalent circuit representations are possible. It is essential, therefore, to establish appropriate criteria on the basis of which it would be possible to determine whether or not a particular coupling structure, i.e., an antenna, is electrically small.

Definitions

D_{max} = maximum dimensions of coupling structure measured from its load terminals to the most distant point on the structure.

f_c = maximum significant frequency component of the EMP spectrum

τ = D_{max}/c -- the time it takes an EM wave to travel the distance D_{max} at the velocity of light c.

Then a coupling structure is electrically small if

$$\boxed{D_{max} \leq \frac{\lambda_{min}}{6}}$$

(Frequency-domain criterion)

or

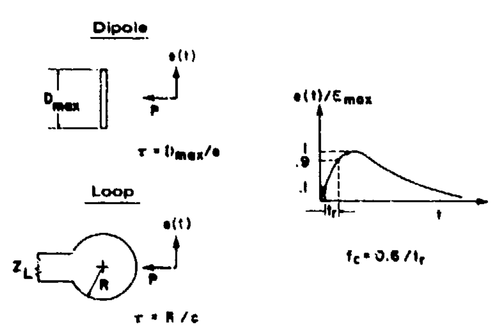
$$\boxed{t_r \geq 4\tau}$$

(Time domain criterion)

where

$f_c = 0.6/t_r$

t_r = rise time of EMP excitation.



Electrically-Small EMP-Coupling Structures

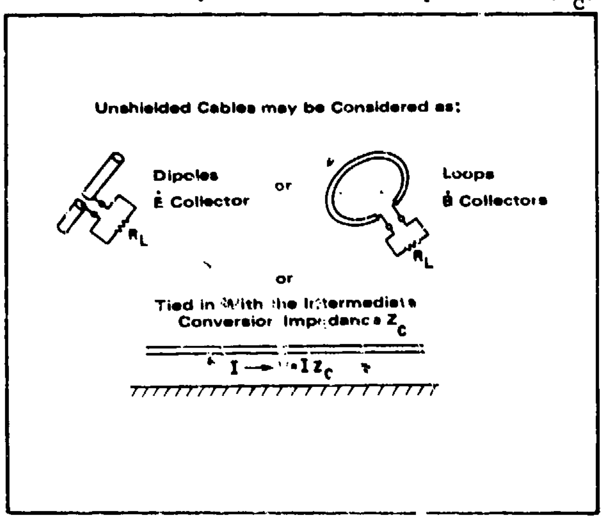
Loop Antennas

Loop type antennas can couple to an electromagnetic wave through either the electric field or the magnetic field. A voltage will be induced in a loop antenna if a time-varying electric field has a component parallel to one of the sides of the loop, or a time-varying magnetic field component normal to the plane of the loop. As stated earlier, a time varying magnetic field induces a current in a loop antenna such that the magnetic field produced by the current opposes the applied field.

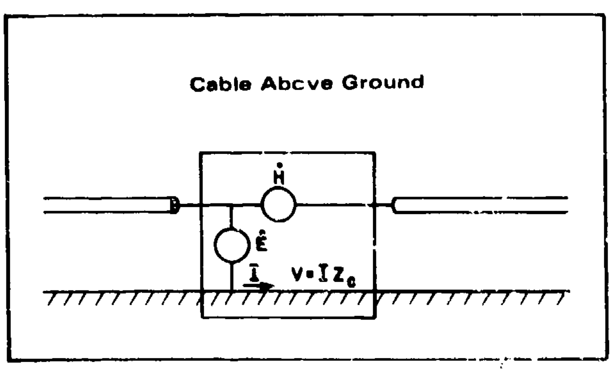
As in the case of linear antennas, a very simple relationship between the field and the induced voltage exists when the loop is electrically small, or stated another way, the field is uniform throughout the area of the loop. For a loop to be considered small, its diameter must be less than $\lambda_{min}/6$. For typical EMP waveforms, this results in a physical diameter of less than 2 feet for a loop in free space.

Cables

Cabling may be considered in two broad categories: (1) unshielded, and (2) shielded. In the case of unshielded cables, they may be considered as $\dot{E}$ and $\dot{B}$ collectors as previously discussed for linear or loop antennas. Coupling to unshielded cables can also occur through an intermediary conversion impedance (Z_c).



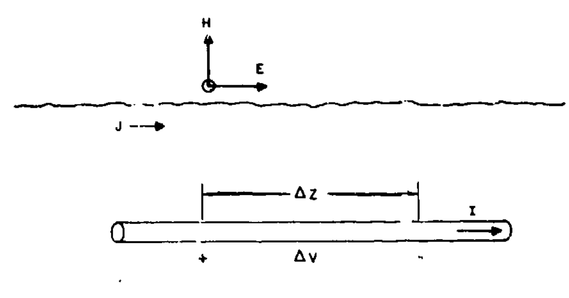
For a cable above, but in proximity to the earth, all three interaction mechanisms are present. The height of the cable above the earth, the earth parameters, and the parameters of the EMP all impact the energy coupled.



The magnetic induction ($\dot{H}$), results from the flux change between the cable and earth. This induced voltage appears as an incremental series voltage source. The electric induction ($\dot{E}$), results from the displacement current which flows between the cable and the earth due to the stray capacitance. It appears an an incremental shunt current source. The voltage source associated with the intermediate conversion impedance (Z_c), results from the current flow in the earth due to its finite conductivity. This source appears as an additional incremental voltage source in series with the load. All of these incremental contributions must be summed in the proper time-phase which takes into account the earth parameters and the angle-of-arrival.

For long runs of cable which run over or within the ground, ground parameters as a function of frequency are important. Also important for the pickup for these long runs is the angle-of-arrival and both the low frequency and the high frequency content of the EMP waveform.

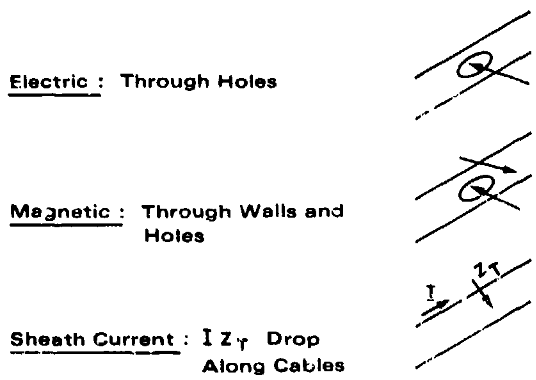
If the cables are buried within the ground, the principal pickup mechanism to cause current to flow on the cables is the common impedance mechanism. Electric and magnetic fields cause currents to flow in the earth, and the resistance of the earth causes a voltage drop to appear along the cable.



BURIED CABLE

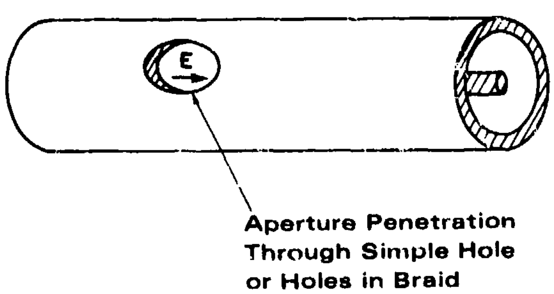
For shielded cables, the shield is viewed as a cylindrical shield. The wires inside the shield interact with the penetrating fields through the same mechanisms ($\vec{E}$ and $\vec{H}$) as the unshielded cables. The penetrating fields, however, are altered by the shield to reduce the field amplitude and change the spectral content. A form of intermediate conversion impedance also plays a role. In the case of shielded cables, it is called the surface transfer impedance (Z_T) and is a characteristic of the cable shield.

Shield Penetration Mechanisms



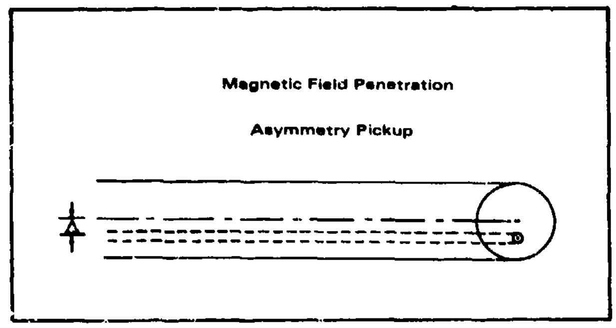
Electric field penetration generally occurs through the apertures or small defects in the exterior shield of the cable. Electric field penetration is also associated with connectors which are not fully shielded. In addition, the electric field is often enhanced at the terminations of long exposed cable runs, and this contributes greatly to electric field penetration effects.

Electric Field Penetration

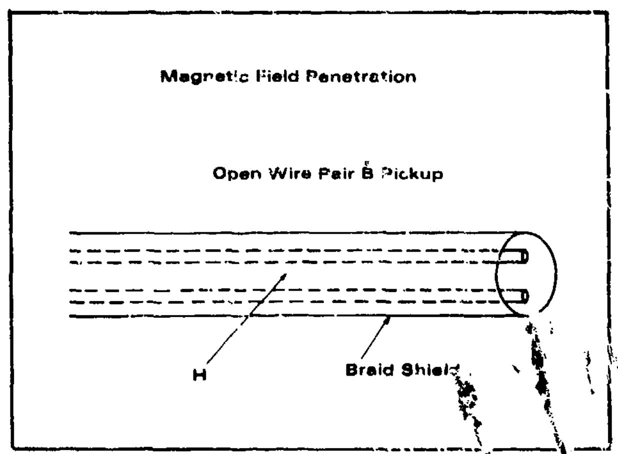


Magnetic field penetration results from aperture coupling and diffusion through the shield, especially for thin, nonferrous shields. The penetration mechanisms for shields is discussed more fully in the shielding section.

In the case of coaxial cables, the interior magnetic field pickup can occur because of the asymmetry, Δ, between the idealized position for the center conductor and the actual position of the center conductor as it occurs in the manufactured cable. This asymmetry essentially results in loops of unequal area and therefore unequal voltage being induced.

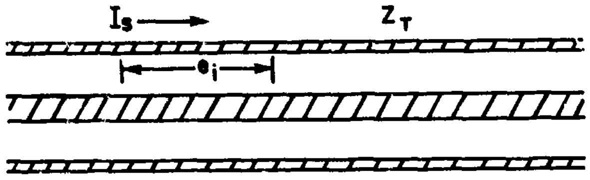


The primary pickup mechanism for shielded open wire pair cables is also through the $\dot{B}$ mechanism. The loop in this instance is closed by the terminations at both ends of the cable.



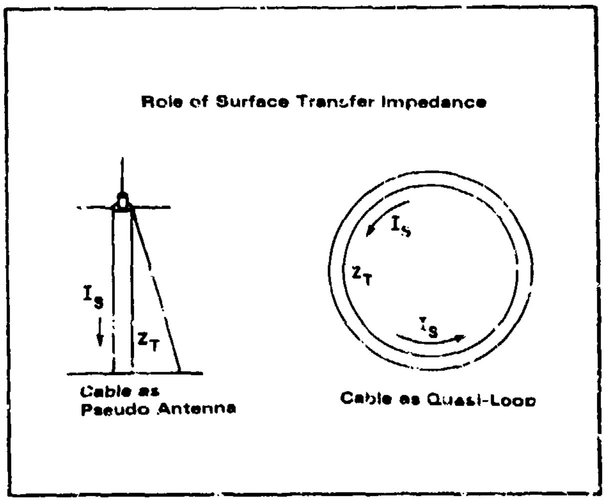
Here we see the definition of the intermediary conversion impedance between the external fields and the inside voltage of a particular cable. The surface transfer impedance is defined as the voltage appearing on the inside of the cable on a per meter basis due to the current flowing on the outside sheath.

Sheath Current Pickup



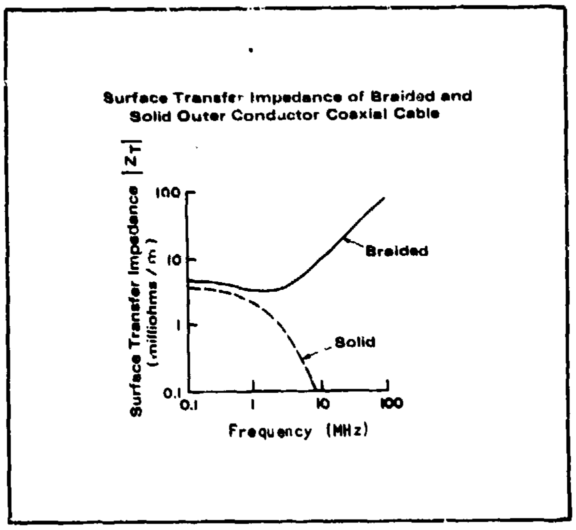
$$Z_T = \frac{e_{\text{inside}}}{I_{\text{sheath}}}$$

The surface transfer impedance plays a very important role in evaluating the cable performance. Transfer impedance is important for electric field pickup because the $\vec{E}$ field can induce currents to flow on cable runs which are parallel to the electric field as illustrated by the cable which terminates in a UHF antenna mast. Sheath current also flows on inadvertent loops formed by cable runs and other metallic structures.



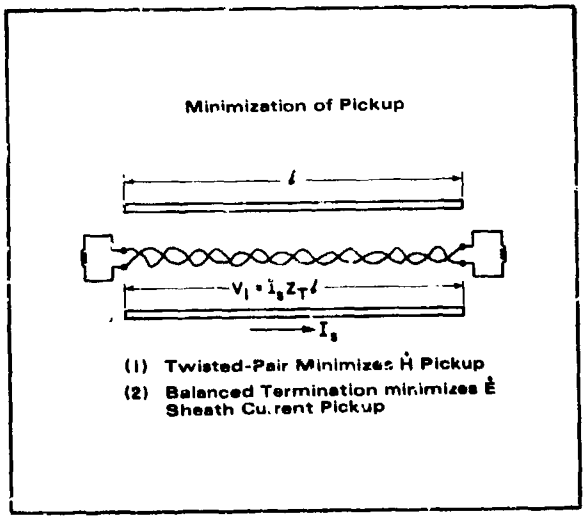
Role of Surface Transfer Impedance

The transfer impedance is very important and at low frequencies is about the same for both braided and solid wall outer shields of the same copper content. However, in the case of the braid, because of the field penetrations through apertures, the transfer impedance increases with frequency, thereby dramatically increasing the pickup characteristics of braided cable. On the other hand, solid wall cable does not exhibit this penetration effect and, therefore, has an improving transfer impedance with frequency primarily because of skin-effect absorption.



Surface Transfer Impedance of Braided and Solid Outer Conductor Coaxial Cable

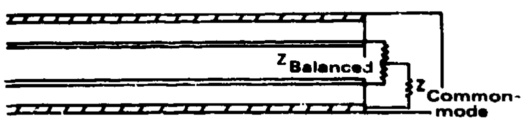
There are some favored approaches regarding cables. One favored approach is the use of twisted wire cable to minimize the $\vec{B}$ pickup. This twisted wire cable is also shielded to minimize the $\vec{E}$ pickup effects. However, the E-field pickups do occur with the presence of sheath currents. The sheath-current-induced inside voltages (common mode pickup) are minimized by the use of balanced terminations which, in effect, are not connected to the external sheath.



Minimization of Pickup

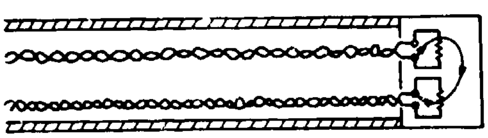
The terminations and splices also can play a major role in enhancing the coupling effects associated with long cable runs. In the case of a balanced pair within a cylindrical shield, the cable should be terminated both for the balance and common mode terminations. Small imperfections along the surface of the cable can convert some of the common mode pickup into differential mode pickup and vice versa. Thus, if the cables are not properly terminated for both modes, the pickup and reverberation effects are greatly enhanced.

Balanced Pair Termination



In the case of multi-conductor bundled cables, all sorts of undesirable effects can be produced by cross-talk effects on the terminations.

Multi-Conductor, Bundled Cable Termination



In summary, cable pickup mechanisms are an extension of $\dot{E}$, $\dot{H}$ and Z_c coupling. The same holds for shielded cables except the effect of the shield must be considered. Improper terminations also play a major role in the pickup associated with cable runs. In the case of very long cable runs, time-delay effects become important and give added weight to angle-of-arrival, earth parameters, and the lower or higher frequency content of the EMP waveform.

Summary for Cable Pickup

- Mechanisms are $\dot{E}$, $\dot{B}$, and Z_c
- Shielding by outer conductor must be considered
- Time delay effects become important
- Terminations can contribute to the pickup

Shielding and Penetration

As discussed previously, propagating EM waves have coupled electric and magnetic fields. In general, whenever you have a time-varying electric field, there is an associated time-varying magnetic field. At low frequencies, however, this coupling is relatively loose and electric and magnetic field penetration effects can be considered separately on a practical basis. This separability of the fields permits defining the figure of merit (shielding effectiveness) of a shield separately for the magnetic and electric field penetration. The inside fields are generally the fields in the geometric center of the enclosure. The outside fields are those fields which would appear at the same spatial point as the inside fields with the enclosure removed.

Shielding Effectiveness at Low Frequencies

$$(SE)_H = 20 \log \left(\frac{H_{inside}}{H_{outside}}\right)$$

$$(SE)_E = 20 \log \left(\frac{E_{inside}}{E_{outside}}\right)$$

At the higher frequencies, the electric and magnetic fields must be considered jointly. In many instances, the power flow is a convenient form of definition. In this case, the power flow, in the absence of the enclosure, is compared to the power flow with the enclosure present. This is somewhat of an unwieldly definition and a more appropriate definition might compare the energy inside the enclosure with the energy in the absence of the enclosure being taken as a stored energy over a specified volume (the volume of the enclosure).

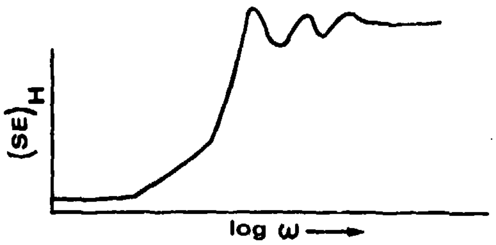
Shielding Effectiveness at High Frequencies

$$(SE)_p = 10 \log \left(\frac{P_{inside}}{P_{outside}} \right)$$

$$(SE)_W = 10 \log \left(\frac{W_{inside}}{W_{outside}} \right)$$

The shielding effectiveness is generally presented as a function of frequency. This is convenient for many radio frequency engineering-type applications but forms a major stumbling block in translating this rather simple concept into the appropriate EMP requirements.

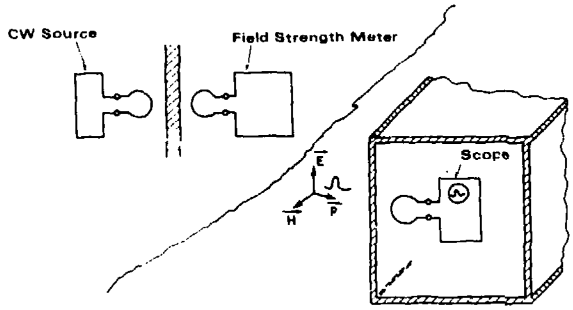
Shielding Effectiveness Presented as a Function of Frequency



It must also be recognized that the shielding effectiveness often stated is a strong function of the measurement method. For example, if the shielding effectiveness measurement is made with two small loops adjacent to the wall, one value of shielding effectiveness is obtained. If, on the other hand, the shielding enclosure is placed within an EMP simulator, another value is obtained. In general, these measures are each self-consistent with frequency but can have values which differ by as much as 20 or 30 dB. The difference between these two measurements is the small loop technique only measures the shielding effectiveness in a local area. The plane wave technique, on the other hand, provides for current flow simultaneously on all surfaces and thus edge and corner effects are seen. The small loop measurements are, therefore, useful to determine seam leakage, etc., and the plane wave for overall structure shielding effectiveness.

In EMP, we are dealing largely with plane waves where the transient response of the enclosure is important. Thus, the amplitude response, as a function of frequency, must be related to the transient response. There are some rather simple ways of doing this which will be considered in more detail. However, before doing this, let us consider on a qualitative basis how shields actually work.

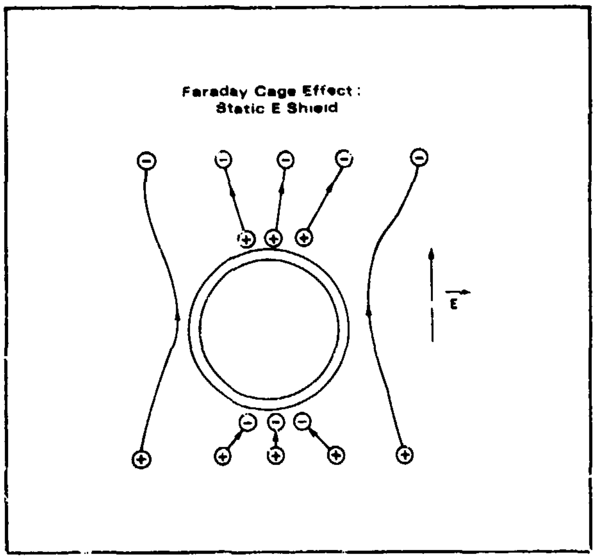
Shielding Effectiveness as a Function of Measurement Method



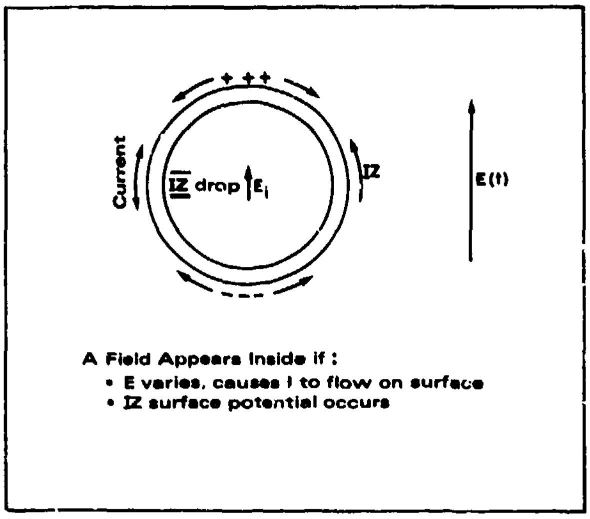
Electric Field Shielding

If a spherical conducting shell is placed within a static electric field, the free charges in the conducting shell will redistribute in accordance with the applied field. This redistribution of charge will be such as to make the force on them zero, or in other words, cancel the applied field. These charges reside only on the surface of the spherical shell and terminate the field lines.

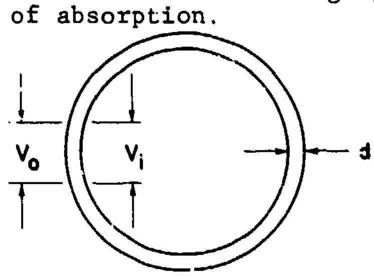
Under these static conditions, once equilibrium has been reached, there is no charge flow and perfect electric field shielding is obtained. This is the well known Faraday Cage effect.



Now allow the applied field to vary slowly with time (a quasi-static field). As in the static case, the free charges will redistribute to reduce the force on them. However, whereas in the static case, equilibrium was reached, in the quasi-static case, the charges will always be in motion since the field is changing amplitude and polarity with time. The result is a time varying current flow on the shell, and since the shell conductivity is finite, a voltage drop results on the surface of the shell. If the shell wall is very thin, the same voltage appears on the inner surface and an electric field is established within the shell.



For very thin-walled enclosures and at low frequencies, the interior field has the same potential as the field on the outside of the enclosure. However, for higher frequencies and thick-wall enclosures, skin-effect mechanisms take place which absorb some of the energy. This causes additional shielding by the mechanism of absorption.



$d << \delta$

At Low Frequencies, $V_{inside} = V_{outside}$

$d >> \delta$

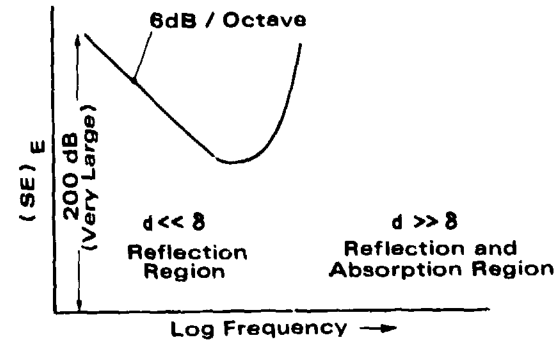
At Higher Frequencies, V_{inside} is Reduced Due to Skin Effect or Absorption

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} = \text{Skin Depth}$$

Typical electric field penetration into an idealized shielded enclosure is shown here. At very low frequencies the charged distribution on the exterior of the enclosure is such as to "cancel" the induced interior fields, giving rise to the Faraday-cage shielding effect. In the case of closed ideal enclosures, this results in almost infinite shielding effectiveness at the very low frequencies. Thus, the electric field is very easy to shield, even with fairly poor conducting materials. On the other hand, if we allow the frequency of interest to increase, we find that the electric shielding effectiveness decreases with an increase in frequency until the absorption losses predominate.

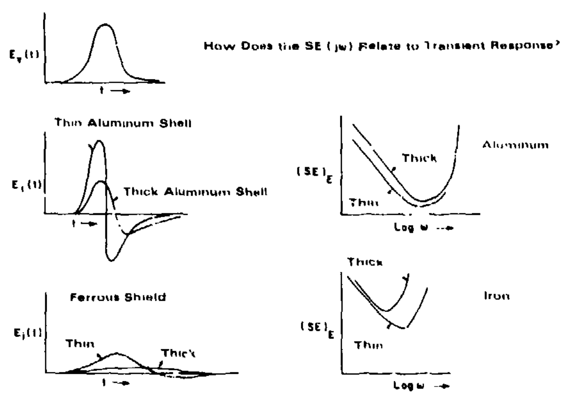
Typical Electric Field

$(SE)_E$ Idealized Shield



Electric Field is Easy to Shield

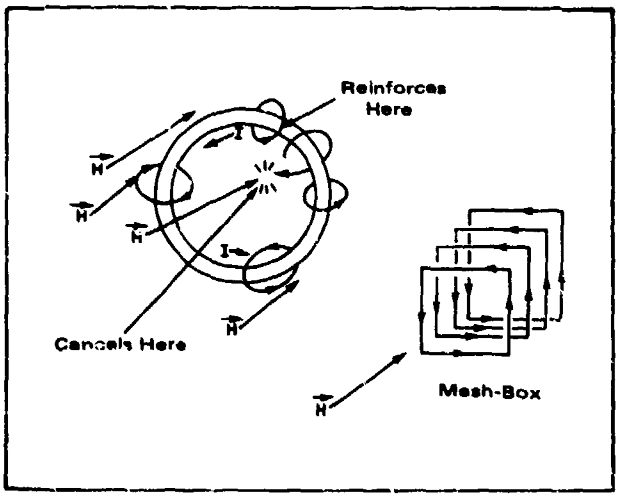
Now consider the relationship between pulsed fields and the shielding effectiveness as a function of frequency. The pulsed field contains a wide spectral distribution. The amplitude of the penetrating field is a function of frequency and so is the phase. For thin wall enclosures, the fields appearing in the interior of the enclosure are roughly the time derivative of the applied (exterior) field. On the other hand, for thick walled enclosures, the amplitude of the interior field is substantially reduced and the derivative effect tends to disappear. This is particularly noticeable in the case of ferrous metal shields due to increased absorption.



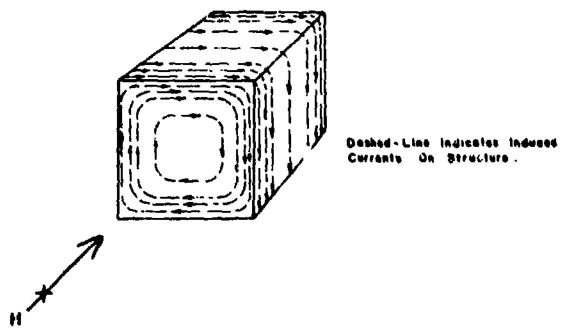
Magnetic Field Shielding

As in the case of electric field shielding, to obtain magnetic field shielding it is necessary to establish an induced magnetic field (interior field) which will cancel the applied field (exterior field). A magnetic field results, in the case of a conductive structure, from the current induced in it. The induced current results from the principle of magnetic induction which requires a time varying applied magnetic field.

Consider a conducting loop in the presence of a time varying magnetic field. The induced current, due to magnetic induction, establishes a magnetic field which reinforces the applied magnetic field exterior to the loop, but opposes or cancels the field in the interior of the loop. If a number of loops were "stacked" to form in the limit, a conducting box or volume shield, the fields on the interior of the box are surpressed at the expense of enhancing the fields outside the box.



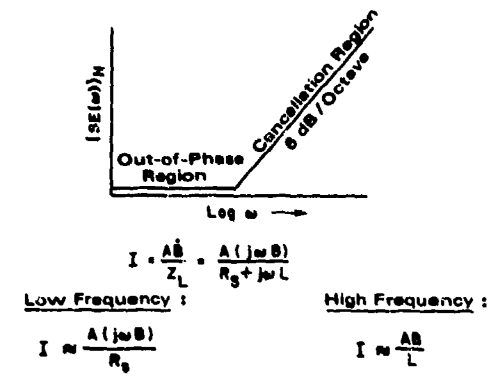
The current flow on a closed box, however, is not uniform. The current flows around the box and near the edges. This occurs because adjacent eddy current elements cancel on the faces normal to the magnetic field.



Induced currents on box from time-varying magnetic field.

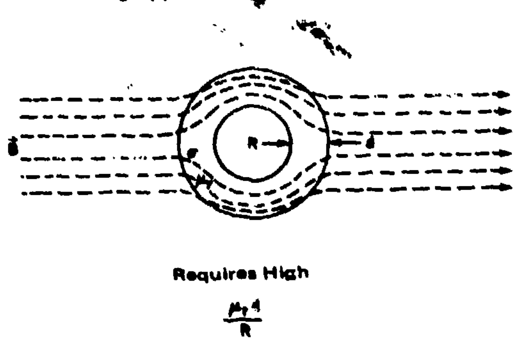
Both the magnitude and phase of the induced current on a conducting box are a function of the frequency of the applied field. The magnetic reflection or cancellation effect occurs because the current flowing in the outer ring becomes more and more in phase with the applied field. At the low frequencies, the current flowing in the outer ring, as seen in the equation, is almost 90° out-of-phase with the applied field. At higher frequencies, current in the outer ring becomes almost in phase with the applied field, and thereby provides better cancellation, which increases the shielding effectiveness at the rate of 6 dB per octave. A necessary condition is that the wavelength is much larger than the radius of the loop.

As the frequency continues to increase, the shielding effectiveness begins increasing at a higher rate. This is due to skin effect absorption as in the electric field case.



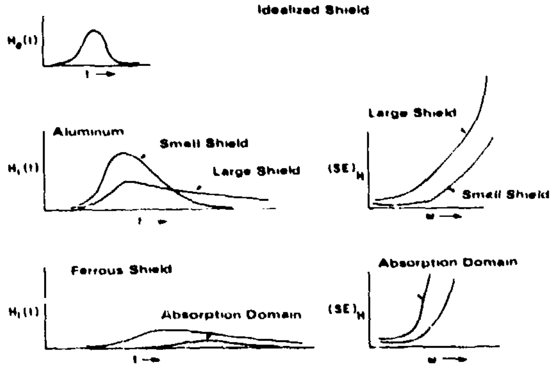
As noted, the shielding effectiveness, as exemplified by ordinary wall materials, such as thin-wall iron, thin-wall sheet iron, copper or aluminum, exhibit very poor low-frequency magnetic shielding effectiveness. Where this is a requirement, the magnetic flux may be ducted away from the area of interest by a ducting "high-perm" shield. This requires, in general, a high product between the relative permeability of the wall material and the ratio of the wall thickness divided by the average radius of the enclosure. In general, such approaches are not practical with ordinary shielding materials because of weight and cost.

Ducting Approach to Low-Frequency Shielding



We have discussed how the magnetic field shielding effectiveness varies with frequency. The relationship of this variation with frequency to a transient response must now be considered. These curves show a typical transient waveform applied to an idealized shield (no apertures or penetrations). Note that the penetrating magnetic field waveform is stretched out by the shield and that a large volume shield exhibits greater magnetic shielding effectiveness than a small volume shield. In the case of ferrous shields, the penetrating waveforms are further reduced due to the higher permeability. The most important feature of the ferrous shield is the stretching out of the rise time of the interior field, resulting in smaller coupling to interior conductors due to magnetic induction (BA) effects. The interior fields are roughly the time integral of the exterior magnetic field and decay exponentially.

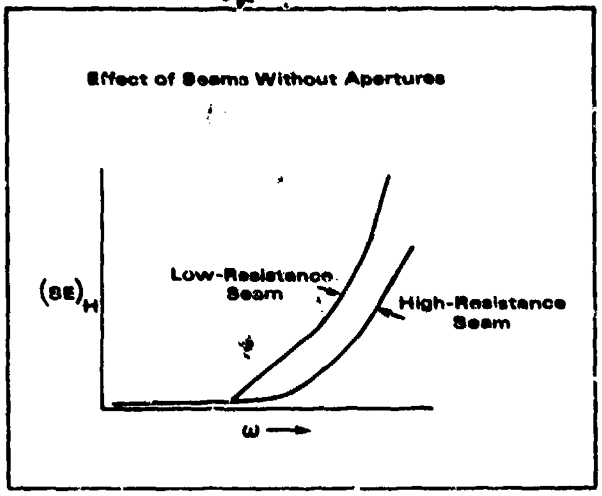
Typical Magnetic Field Shielding Effectiveness
Idealized Shield



Seams and Apertures

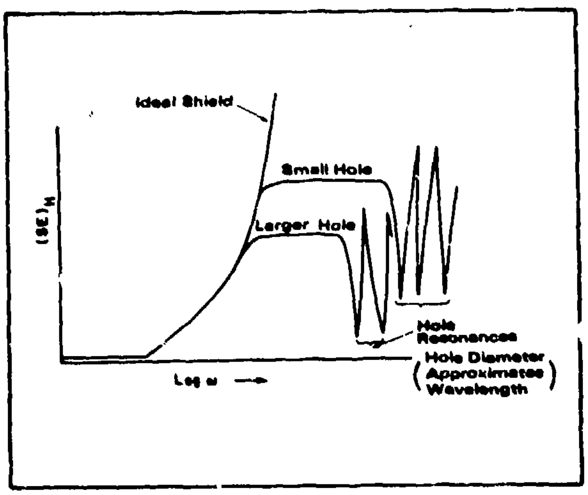
The previous discussion of shielding effectiveness was concerned with idealized shielding structures containing no seams or apertures. In any practical enclosure, these idealized conditions cannot be met. The primary effect of seams and apertures is to reduce the overall shielding effectiveness.

In general, enclosures which have high resistance (as compared to the enclosure material), result in poorer shielding performance than those with low resistance seams. This degradation in performance results from the added series resistance reducing the current flow on the enclosure. This reduced current flow reduces the bucking magnetic field and, consequently, the magnetic shielding effectiveness. As we will see later, it also produces a secondary effect of introducing a voltage drop across the seam which appears on the inside of the enclosure which is very important if multipoint grounds are utilized.

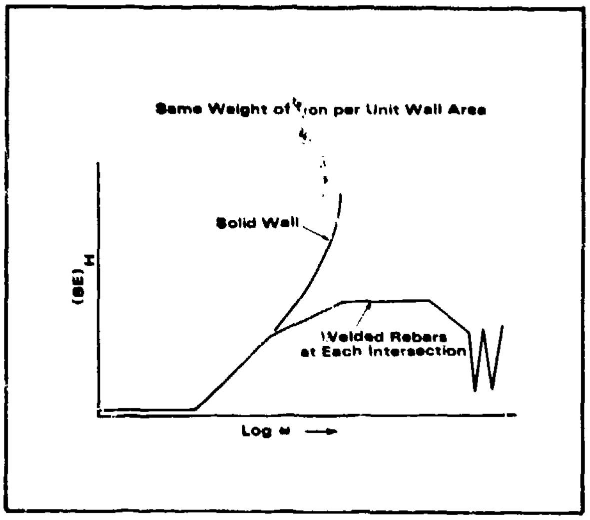


Obviously, if the enclosure has openings, this will provide a means for exterior fields to leak into the interior. Since these openings (assuming the opening to be small relative to the size of the enclosure, can be viewed as aperture or slot antennas, the fields at low frequencies (below resonance) tend to fall off inversely proportional to the cube of the distance from the aperture.

The effect of these apertures is most significant at the higher frequencies and essentially limits the high frequency performance of an enclosure. As frequency increases, the fields penetrate to the interior, the frequency at which this begins being dependent on the hole size. As the frequency continues to increase, resonant penetration can occur as indicated by the wide swings (very low attenuation of field) in the shielding effectiveness curve.



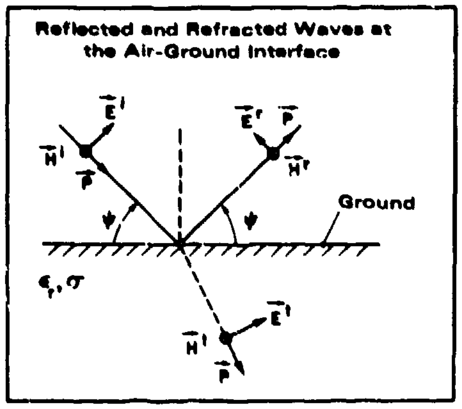
This shows a comparison between the idealized solid-wall enclosure with that of a welded rebar enclosure with about the same enclosure weight. Note that much greater shielding effectiveness performance is realized by the solid wall enclosure. At the higher frequencies, the shielding effectiveness of the welded rebar enclosure tends to level off for a variety of reasons, such as skin-effect, and finally becomes ineffective due to resonant penetration through the apertures.



Ground Effects

Up to this point we have stressed the interaction of a plane propagating EM wave with a system. In the case of systems which are isolated from the earth or other structures, such as aircraft or satellites, this is a good approximation. However, most of our systems are located near, on, or under the earth's surface. For this majority of systems the effects of the earth on the EM field must be considered.

At any interface between two media of differing characteristics, an EM wave will undergo reflection, refraction, and absorption. The percentage of the wave reflected, or transmitted, is a function of both the wave characteristics and the media characteristics. For purposes of discussion, we will assume an air/earth interface. In this instance, the characteristics for the propagation in air are the same as free space. The most important parameters of the incident EM wave are its spectral content, angle of incidence, and polarization, since the reflection coefficient (or conversely, the transmission coefficient) are strong functions of these parameters. It should be noted that phase reversal occurs for the component of the electric field parallel to a conducting surface, but no phase reversal for the component normal to the conducting surface. We can see, therefore, that the angle of incidence and polarization determine the components of the electric field with respect to the reflecting surface and consequently, the reflection or transmission coefficient.



The media (earth) parameters of interest are its conductivity (σ), relative permittivity (ϵ_r), and relative permeability (μ_r). Typical values for these parameters for earth are:

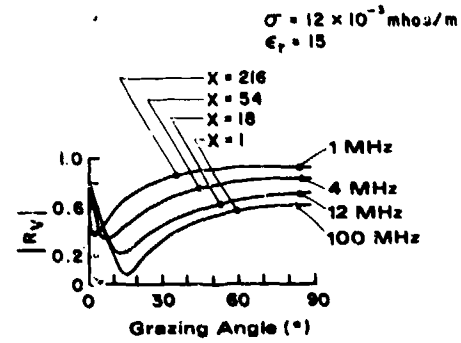
$$\sigma = 10^{-2} \text{ to } 10^{-4} \text{ mhos/m,}$$

$$\epsilon_r = 10 \text{ to } 15, \text{ and}$$

$$\mu_r = 1.$$

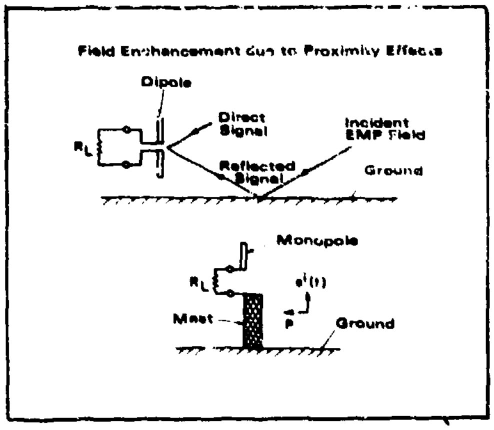
In general, the higher the conductivity, the less the penetration of the wave into the earth. A plot of the magnitude of the reflection coefficient for vertical polarization is shown in the figure as a function of the angle of incidence and frequency. As the angle of incidence increases, there is a sharp dip in the curve (which corresponds to the Brewster angle in optics) where maximum transmission (minimum reflection) is realized. The width of this dip is quite narrow in terms of the grazing angle.

Magnitude of the Plane Wave Reflection Coefficient for Vertical Polarization



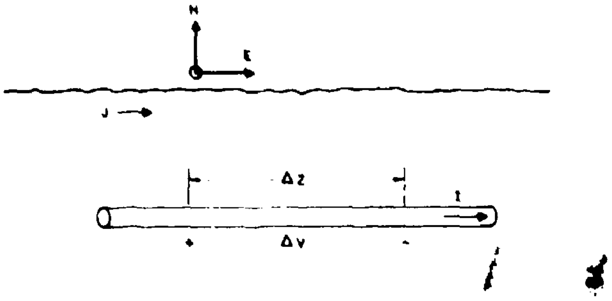
The importance of this earth reflection is the EM fields incident on systems in proximity to the earth are considerably different than the free space condition. For buried systems, the transmission coefficient is of prime concern. The high frequencies will also be absorbed more (higher attenuation factor, than the low frequencies. Therefore, for buried systems, the EM spectra will be predominantly low frequency and of reduced amplitude if the system is buried more than a few meters (>10 m) below the surface.

For systems on or near the ground, the magnitude and phase of the energy reflected are of prime concern. The EM fields incident on the system of interest will be the vector sum of the incident (free space) and reflected waves. This can result in either enhancement or reduction of the fields impinging on the system. One such case is a dipole above ground as shown in the figure.



Another case of interest is placing an antenna on a mast removing it from proximity to ground. This also can significantly alter the EM energy collection of the system. If the mast is a conducting structure, the monopole and mast configuration can be viewed as an asymmetrical dipole in which the mast contributes substantially to the induced current on the structure.

If cables are buried within the earth, the principal pickup mechanism to cause sheath current to flow on the cables is the common impedance mechanism. Electric and magnetic fields cause currents to flow in the earth, and the resistance of the earth causes a voltage drop to appear along the cable.



BURIED CABLE

4.3 COUPLING AND INTERACTION ANALYSIS

The preceding discussion has considered, on a qualitative basis, how EM waves interact with and couple to systems. In this section, we will discuss briefly what role analysis plays in assessing the EMP vulnerability of a system, how one can obtain a system model for analysis, and some of the mathematical tools employed in analysis.

Role of Analysis

The role analysis may play in a vulnerability assessment can be quite varied. During system design, no hardware may be available for test. Analysis to predict the system response, identify weak areas, and provide design assurance may be the only alternative. Analysis during the design, or even after the hardware implementation, is useful to guide the test program or confirm the test results. Since, as we will see later, testing only provides specific answers (it is not practical to try to test for all threats), analysis is useful to extend the test results.

The Role of Analysis

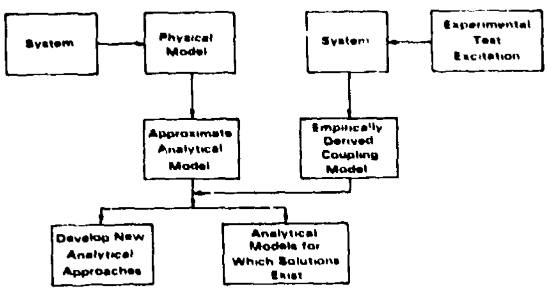
- Predict system response
- Identify weak points
- Provide design assurance
- Guide the test program
- Confirm test results
- Extend test results

Therefore, analysis is not the complete solution any more than testing is a complete solution. Each has a role and both should be used in various combinations to cost effectively predict or evaluate system hardness.

System Modeling

The first step in analysis is the development of a model which is a mathematical representation of a physical system. In general, actual system geometries are too complex to be amenable to analysis. Therefore, it is necessary to develop a physical model (coupling model) of the system. The physical model is one for which solutions of the field equations exist from antenna or transmission line theory. Usually these models are surfaces of revolution such as cylinders and spheres, or for more complex structures, combinations of intersecting surfaces of revolution or wire models. These bodies can be analyzed to determine surface currents using available antenna theory which will be discussed later. An alternative way of obtaining the coupling transfer functions is to subject the system to a test excitation. This provides empirically derived waveforms for the current distributions which can then be mathematically modeled.

Modeling a System



There are a variety of analytical techniques for determining the EM coupling by antennas or structures that can be modeled as antennas. The more widely used of these techniques include: (1) Fourier Transformation Method (FTM), (2) Lumped Parameter Network Method (LPN), (3) Singularity Expansion Method (SEM), and (4) a technique based on Landt's Method.

The Fourier Transformation Method utilizes Fourier Analysis to determine the antenna parameters, effective length and impedance, as a function of frequency to define a Thevenin equivalent circuit for the antenna. The load must be linear (i.e., not a function of the driving voltage) to utilize this technique. The Thevenin Circuit is solved for the current in the load as a function of frequency. The load current is transformed back into the time domain by the inverse Fourier transform.

The lumped parameter network method provides a technique for the analysis of a distributed parameter network (the antenna) connected to a linear or nonlinear load. Like the FTM method, it utilizes a Thevenin equivalent circuit to represent the antenna in times of its open circuit voltages. Both antenna parameters, the effective length and impedance, are then approximated by national network functions that are realizable in terms of an RLC circuit. After the lumped parameter network representation is obtained, stand and circuit analysis computer codes may be used to calculate the transient system response.

In classical circuit theory, the time domain solution of a linear circuit excited by an exterior waveform may be determined from the knowledge of the location of any singularities of the transfer function and its corresponding residues. The transient behavior of the circuit is then obtained as the sum of damped sinusoids whose coefficients are determined by the Residue Theorem. The singularity expansion method has extended this concept to solve electromagnetic boundary value problems.

Basically, this method involves the determination of the antenna response in terms of singularities in the complex frequency domain which represent the natural frequencies, modes, and coupling factors. The antenna response in the time domain is obtained by taking the inverse transform of the terms in the singularity expansion.

The technique based on Landt's method is to solve for the peak short circuit current in an infinite wire antenna. This method utilizes the impulse response of the antenna. Since the peak current on a wire antenna excited by an EMP usually occurs during the early time response of the structure, the impulse response of the current on an infinite wire antenna is valid on a finite wire antenna, up until the time that the reflection is seen.

Analytical Techniques

- Fourier Transformation Method (FTM)
- Lumped Parameter Network Method (LPN)
- Singularity Expansion Method (SEM)
- Landt's Method

The first three methods are the more rigorous methods and require the use of numerical or digital techniques for their solution. The computer codes for computing antenna and cable response to an EMP are maintained for customer use by the Electromagnetic and Systems Research Group, Lawrence Livermore Laboratory, Livermore, CA. 94550. Maintenance of this computer code library is an ongoing effort funded by the Defense Nuclear Agency.

Generally, the analysis of system response to electromagnetic phenomena is divided into two classes: analysis of linear systems, and the analysis of nonlinear systems. We will consider the applicability of the analytical techniques to these two problem areas.

4.4 ANTENNA COUPLING ANALYSIS - LINEAR SYSTEMS

If a system is linear, or can be approximated to be so, Fourier Transform and ac circuit analysis techniques can be used.

Linear Systems

Antennas

- **Linear**
- **Passive**
- **Distributed**

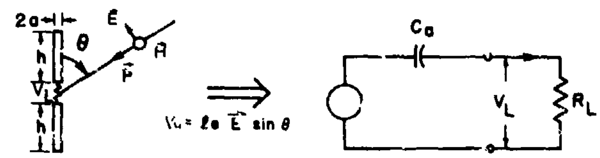
Antenna Loads

- **Linear**

Simple Energy Collectors

The simplest approach to analyzing the antenna coupling in linear systems is where the antenna (structure) dimensions are small for all wavelengths. Consider a short dipole antenna. The equivalent circuit for a short dipole is a voltage source a series capacitor (antenna capacitance), and the load.

Equivalent Circuit of a Small Dipole



$$V_a = \frac{1}{C_a}\int i\, dt + i R_L$$

The response of this circuit can be obtained by inspection for the following loads:

HIGH-IMPEDANCE RESISTANCE LOAD

$$R_L >> 1/\omega C_a$$

then

$$V_L(t) = -h \sin\theta\, e^i(t)$$

LOW-IMPEDANCE RESISTANCE LOAD

$$R_L << 1/\omega C_a$$

then

$$i_L(t) = -h \sin\theta\, C_a \dot{e}i(t)$$

The antenna capacitance, C_a, is given by:

$$C_a = h/cZ_o$$

where

Z_o = 60 (Ω-2) [average antenna characteristic impedance]

Ω = $2\ell n(2h/a)$ [antenna shape factor]

c = velocity of light

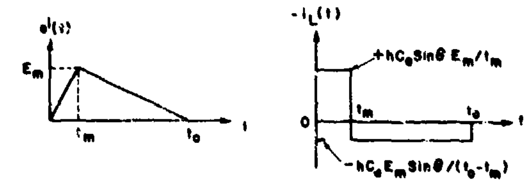
a = antenna radius

h = ℓ_e

The time domain response is approximately the derivative of the incident electric field for the case $R_L << 1/\omega C_a$.

Response of a Small Dipole Antenna to EMP

$R_L << 1/\omega C_a$



Approximate EMP Excitation — Dipole Response

Next, consider a small magnetic dipole (loop) in the presence of an incident EMP field. In the equivalent circuit, L is the low-frequency inductance of the loop and is given by:

$$L = \mu_o R \, [\ln (8R/a) - 2]$$

for $R >> a$

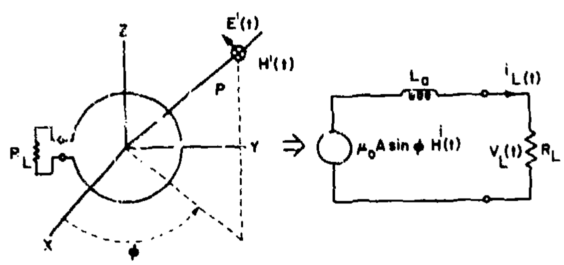
where

μ_o = permeability of free space

R = radius of loop

a = antenna wire radius

Equivalent Circuit of a Small Loop



$$Va = L \frac{di}{dt} + iR_L$$

The response of the equivalent circuit is readily obtainable for the following loads:

HIGH-RESISTANCE LOAD

$R_L >> \omega L$

then

$$V_L(t) = \mu_o A \sin\phi \, \dot{H}^i(t)$$

LOW-RESISTANCE LOAD

$R_L << \omega L$

$$i_L(t) = \frac{\mu_o A}{L} \sin\phi \, H^i(t)$$

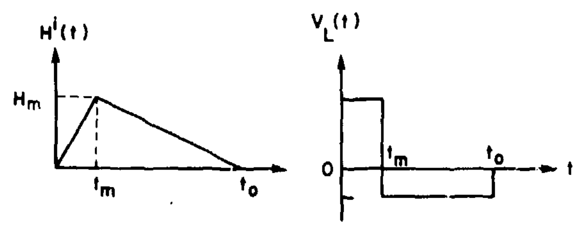
or

$$i_L(t) = \frac{\mu_o A}{\eta_o L} \sin\phi \, E^i(t)$$

Note that the load impedance does not have to be resistive and it may represent the input impedance of an electronic system such as, for example, a receiver front end. The time domain response for the case of $R_L >> \omega L$ is the derivative of the applied magnetic field. For the small dipole, the open circuit voltage ($R_L >> 1/\omega C_a$) follows the EMP waveform in early time. For the small loop, the short circuit current follows the EMP waveform.

Response of a Small Loop to EMP

$R_L >> \omega L$



Approximate EMP Excitation — Loop Response

A quick look approximate energy analysis can also be performed to determine an estimate of the energy dissipated in the load. In this case, the concept of antenna effective area can be employed. The effective area is related to the power gain by:

$$A_e(\omega) = \frac{\lambda^2 G}{4\pi}$$

where

$A_e(\omega)$ = effective area as a function of frequency

λ = wavelength of wave

G = antenna power gain.

The effective area for various antennas is given below:

EFFECTIVE AREA OF VARIOUS ANTENNAS

Antenna Type	Effective Area
Isotrophic Radiator	$\lambda^2/4\pi$
Very Short Dipole	$3\lambda^2/8\pi$
Half-Wave Dipole	$1.64\lambda^2/4\pi$
Large Aperture Antenna	100% physical area
Pyramidal Horn	50% physical area
Parabolic Reflector	50-55% physical area

The solution for an EMP is obtained by calculating the energy available at the antenna terminals in the frequency domain for each significant frequency component in the pulse. The total energy delivered to the load is then the integral (sum) over the frequency spectrum of the pulse which can be determined through normal Fourier analysis techniques.

For an antenna placed in the field of a linearly polarized EM wave, the power available at the antenna terminals under conjugate matched conditions for sinusoidal fields is given by:

$$W = PA_e$$

where

W = available power (watts)

P = power density (watts/m^2)

A_e = effective area (m^2)

For an EM pulse, the energy available at the antenna terminal is:

$$J_T = \frac{1}{2\pi} \int_{-\infty}^{\infty} A_e(\omega)\, J(\omega)\, d\omega$$

where

J_T = available energy (joules)

$A_e(\omega)$ = effective area (m^2)

$J(\omega)$ = energy density spectrum of the pulse (joules/m^2 Hz)

It should be noted that all the available energy will be transferred to the load only when the load impedance presents a conjugate match to the antenna impedance over the frequency range where the excitation has significant components. Such a wide band match is physically unrealizable and would result in too much of a worst case. To accurately calculate the load energy, one would need to know the system transfer function. An approximation can be obtained by making appropriate assumptions.

Assume a transfer function, $T(\omega)$, of the form:

$$T(\omega) = \begin{cases} A_e(\omega_o)\left[\dfrac{\omega}{\omega_o-\omega_1}\right]^{2n} & 0 \le \omega \le (\omega_o-\omega_1) \\ A_e(\omega_o) & (\omega_o-\omega_1) \le \omega \le (\omega_o+\omega_1) \\ A_e(\omega_o)\left[\dfrac{\omega_o+\omega_1}{\omega}\right]^{2n} & (\omega_o+\omega_1) \le \omega \end{cases}$$

ω_o = center frequency of the system response

$(\omega_o-\omega_1)$ = lower cutoff frequency of the system response

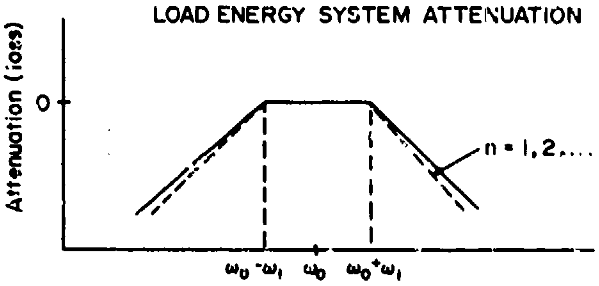
$(\omega_o+\omega_1)$ = upper cutoff frequency of the system response

and n determines out-of-band attenuation.

Under this assumption, the energy dissipated in the load is:

$$J_T = \frac{1}{2\pi} \int_{-\infty}^{\infty} T(\omega)\ J(\omega)\ d\omega$$

The energy available at the load will be distributed as shown below:



For ideal preselection, the rectangle formed by $(\omega_o-\omega_1)$ and $(\omega_o+\omega_1)$ dotted lines would be obtained. The case of n = 1 (6 dB/octave) is considered a worst case and the bandwidth chosen is the turning range of the system. If additional information is available, a different value of n may be more appropriate.

Fourier Transform Method

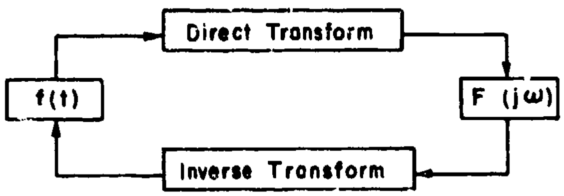
The response of a linear system to pulse or transient excitation can be determined by the use of Fourier Transforms (FT), provided the complex transfer functions are known over the frequency range where the excitation has significant components.

The Fourier transform pair is defined in terms of the following integral expressions:

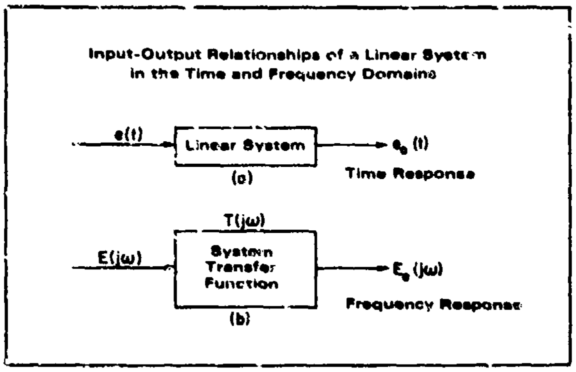
$$F(j\omega) = \int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt \quad \text{[direct transform]}$$

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F(j\omega)e^{j\omega t}d\omega \quad \text{[inverse transform]}$$

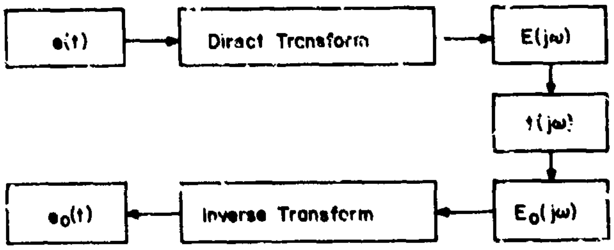
In effect, the direct Fourier transform takes a function from the time domain to the frequency domain, whereas the inverse Fourier transform performs a frequency-to-time domain transformation. Thus, the Fourier transform pair provides a two-way transformation. Note that $F(\omega)$ is the frequency spectrum of f(t).



In principle, the solution to the problem of determining the response of a linear system can be obtained by Fourier analysis as indicated. Note that the concept of system transfer function is valid only for linear systems.



This is a flow chart indicating the procedure for obtaining the time response of a linear system to any excitation using Fourier analysis. In most cases, the Fourier integrations would have to be carried out using numerical techniques with the aid of a digital computer.



In general, the voltage transfer function is

$$T_V(\omega) = \frac{V_L(\omega)}{E_i(\omega)} \quad \text{(volts/Hz/(volts/m-Hz)}$$

and the current transfer function is

$$T_I(\omega) = \frac{I_L(\omega)}{E_i(\omega)} \quad \text{(amps/Hz/(volts/m-Hz)}$$

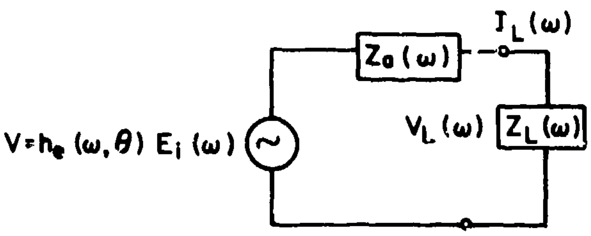
The time dependence is then determined by

$$V_L(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} T_V(\omega) \quad E_i(\omega) \quad e^{j\omega t} d\omega$$

(volts)

and similarly for the current.

As an example of the FTM, the equivalent circuit for an antenna is shown in the following figure



where

$h_e(\omega,\theta)$ = complex effective length of the antenna as a function of polar angle θ and frequency

$E_i(\omega)$ = frequency domain representation of the incident field

$Z_a(\omega)$ = antenna impedance as a function of frequency

$Z_L(\omega)$ = load impedance as a function of frequency

$I_L(\omega)$ = load current as a function of frequency

$V_L(\omega)$ = load voltage as a function of frequency

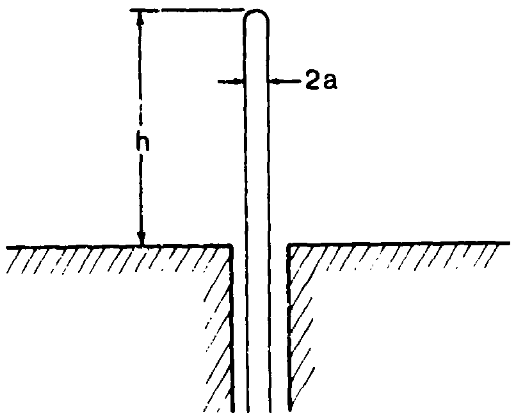
The cases which will be considered will be for monopole antennas with the following load conditions:

Z_L = 50 Ω

Z_L = 0 (short circuit)

Z_L = ∞ (open circuit)

The antenna configuration is shown in the following figure:



CYLINDRICAL MONOPOLE ANTENNA

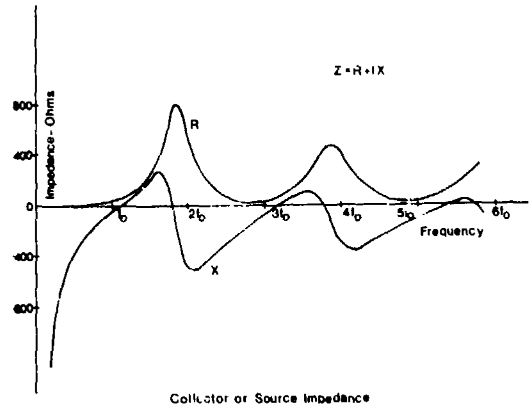
The antenna impedance is a complex function. For

$$\beta h = \frac{2\pi h}{\lambda} \leq 1.0$$

where

$$\beta = \frac{2\pi}{\lambda}$$

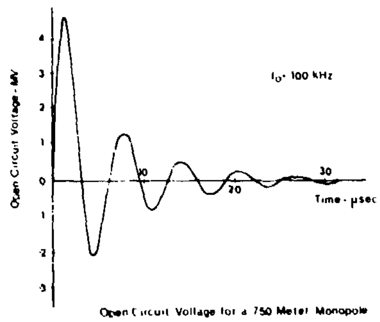
the theory of R.W.P. King, C.W. Harrison, Jr., and D.H. Dentor, Jr., can be applied. For $\beta h > 1.0$, Wu's formulae apply. The antenna impedance for a representative antenna is shown in the following figure.



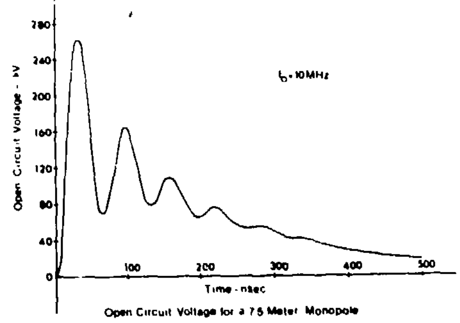
Collector or Source Impedance

Assuming a typical incident high altitude EMP waveform, the time histories of the antenna load voltage and current were calculated using Fourier Transform Techniques.

The time history for the load voltage for $Z_L = \infty$, and for antenna lengths of 750 meters (f_o = 100 kHz), and 7.5 meters (f_o = 10 MHz) are shown below.



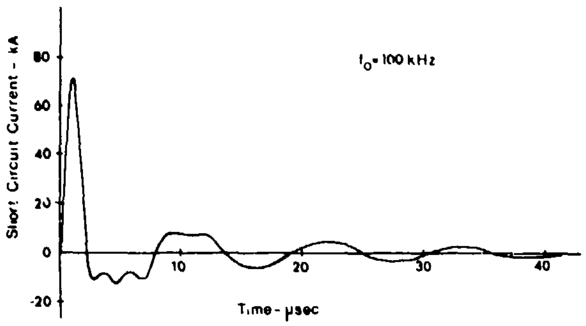
Open Circuit Voltage for a 750 Meter Monopole



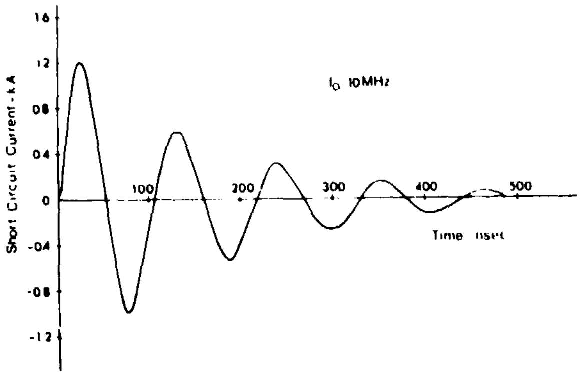
Open Circuit Voltage for a 7.5 Meter Monopole

It should be noted that these antennas basically ring at twice their resonant frequency (damped sinusoid) under matched load conditions. The current distribution approximates that of a shorted dipole in free space under open circuit conditions. In the case of the 7.5 meter antenna, a capacity effect can be seen. This is due to the fact that the antenna is electrically short over most of the frequency spectrum and acts as a capacitance.

The load current for the case of $Z_L = 0$ is shown for these antennas in the following figures.

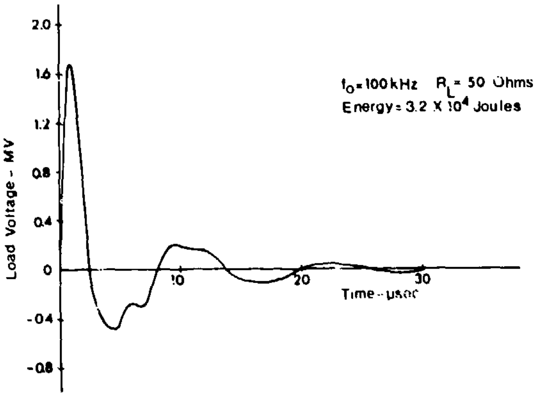


Short Circuit Current for a 750 Meter Monopole

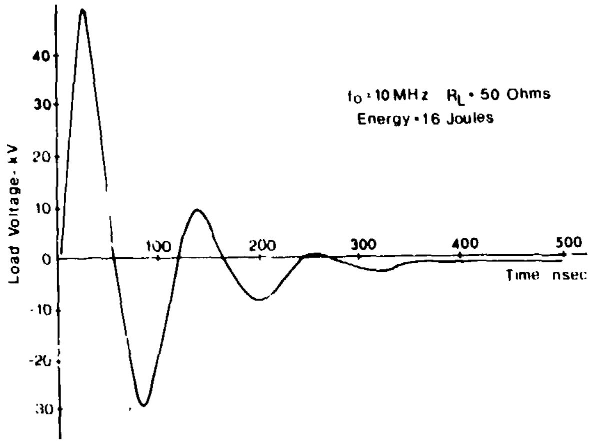


Short Circuit Current for a 7.5 Meter Monopole

The load voltage for the case of $Z_L = 50\ \Omega$ is shown in the following figures.



Load Voltage for a 750 Meter Monopole

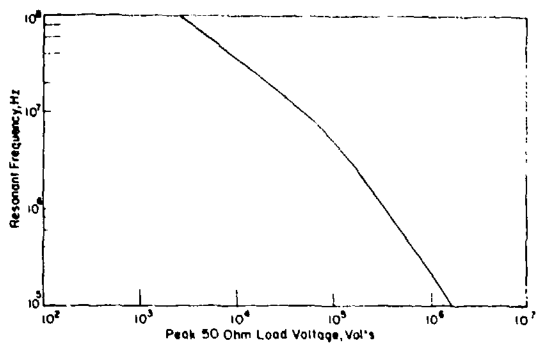


Load Voltage for a 7.5 Meter Monopole

Again, the antenna ringing at the resonant frequency is noted.

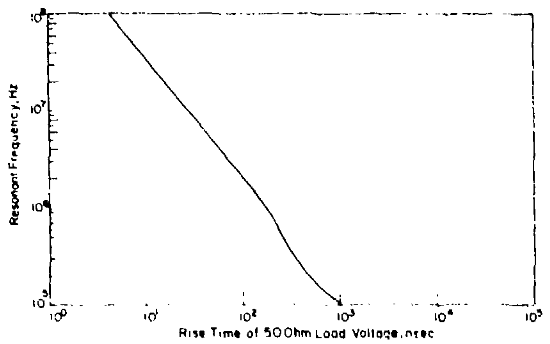
This information can be presented in a more useful manner for system hardening. Of primary interest are the peak voltages, rise times, rate of rise, decay time, and energy for the case of $Z_L = 50\ \Omega$.

The peak voltage for the case $Z_L = 50\ \Omega$ as a function of the resonant frequency of the antenna is shown in the following figure.



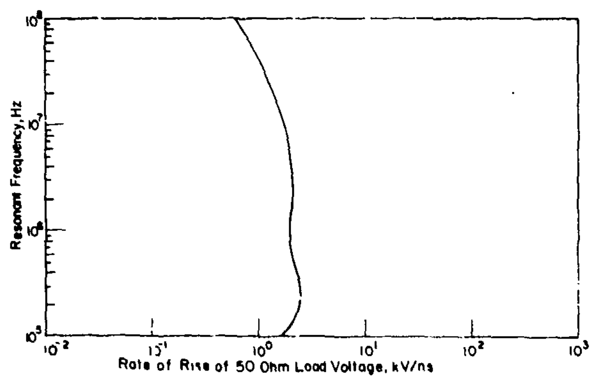
As can be seen, the peak voltage is a direct function of the resonant frequency (antenna length).

The rise time (defined as the time for the amplitude of the initial cycle to increase from 10% - 90% of the peak value) is also of interest for hardening design. This is shown for the case of $Z_L = 50\ \Omega$ in the following figure.



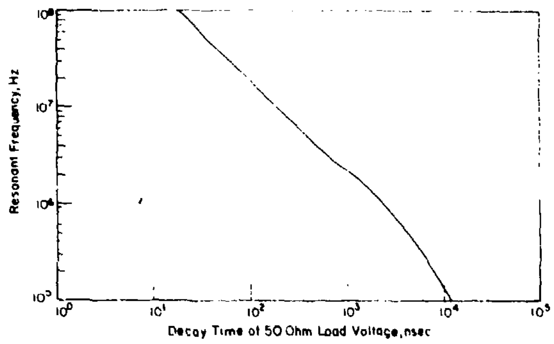
As can be seen, the total rise time is also a direct function of the antenna length as one would expect, since the wave is a resonant ringing damped sinusoid.

Surge protection devices (Section VI), however, are primarily sensitive to the rate of rise of the voltage. The rate of rise as a function of antenna length (resonant frequency) is shown in the following figure.



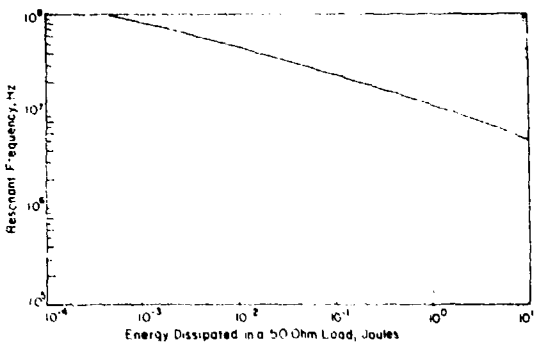
As can be seen, the rate of rise (kV/ns) is relatively independent of the antenna length. Consequently, surge arrestors must have essentially the same response characteristics regardless of the resonant frequency of the antenna.

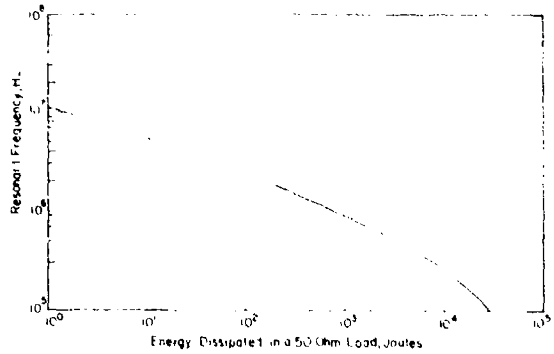
Of equal importance is the decay time of the pulse in order to determine the energy coupled into the system. This is shown in the following figure for $Z_L = 50\ \Omega$.



Again, the decay time is a direct function of the antenna length.

The total energy dissipated in the load, $Z_L = 50\ \Omega$, is directly related to antenna length (resonant frequency). This is shown in the following figures.



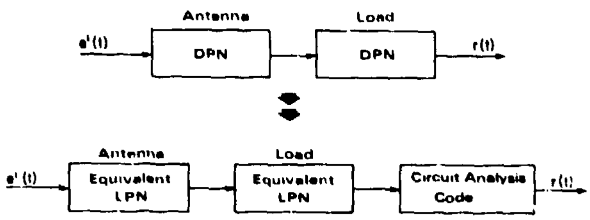


4.5 ANTENNA COUPLING ANALYSIS - NONLINEAR SYSTEMS

If a given system is nonlinear, Fourier Transform methods are not applicable and one must solve the resultant differential equations describing the behavior of the system. Sometimes standard circuit analysis computer codes, such as SCEPTRE, are suitable for this purpose, if the system can be characterized in terms of lumped parameters, i.e., resistors, capacitors, inductors, and controlled sources. Most electronic systems are, in fact, nonlinear since they contain such devices as diodes, tubes, transistors, etc. The major difficulty is deriving this lumped parameter network. The network must have the same transient response as the distributed system it is to represent. Both analytical and experimental techniques have been utilized to obtain the system transient response. Having the transient response, standard circuit synthesis approaches can be used to define the LPN which would produce this response.

In such cases, standard circuit analysis computer codes, such as SCEPTRE, CIRCUS, or others can be employed. This creates the need for a lumped-parameter network (LPN) representation of an antenna, which is basically a distributed network.

Response of a Nonlinear System using the Lumped-Parameter Network Method



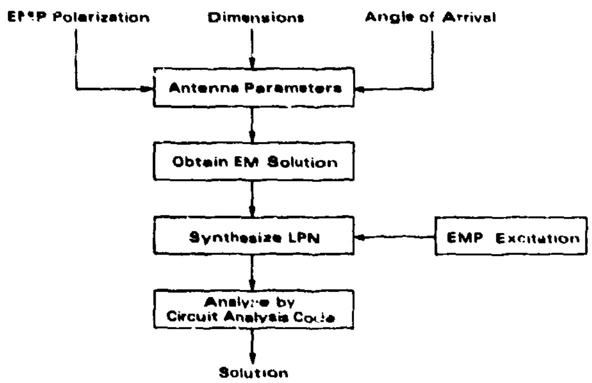
It is to be noted that:

1) Frequency domain techniques are not valid when nonlinear elements are incorporated into a system.

2) If the lumped-parameter network (LPN) representation of a system is given standard analysis computer codes can be used for a nonlinear system.

A flow chart of the lumped-parameter network method for obtaining the transient response of an antenna system would be:

Flow Chart of Lumped-Parameter Network Method



To obtain the lumped parameter network (LPN) equivalents of the effective lengths and impedances for monopole and dipole antennas, the expressions given by Wu were analytically continued over the complex frequency plane, and the poles and zeros of these functions were found numerically. The Singularity Expansion Method (SEM) is one alternative way of determining the required poles and zeros. From a truncated set of poles and zeros, the functions resulting were optimized over finite frequency ranges. Using these functions, conventional network synthesis procedures were used to find the network element values.

Another alternative approach to determining the effective length and impedence for an antenna would be experimentally. The experimental data could be fit by again using conventional synthesis approaches.

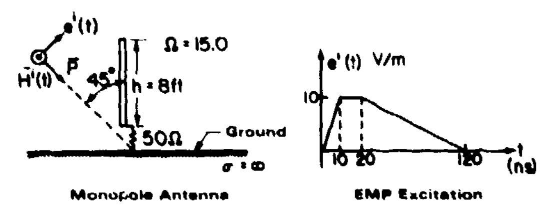
At the present time, a catalog of equivalent circuits only exists for monopoles and dipoles. Effort is continuing to develop equivalent circuits for more complex antennas.

Availability of LPN Representations of Antennas

- Complete tables of element values are available only for monopole, dipole and folded dipole antennas. These tables take into account antenna dimensions and any angle of arrival.
- Tables of element values for other antenna configurations are presently under development.
- Tables of element values are also available for the ground reflection coefficient. Such tables are used in calculating the transient response of antennas in proximity to ground.

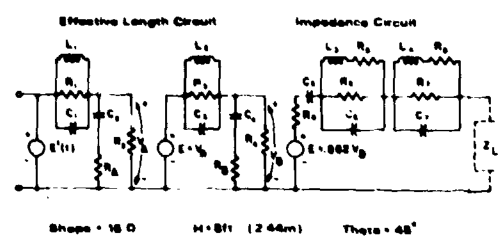
As an example of the use of the LPN method, the transient response of an 8 foot monopole over a perfectly conducting ground was computed using both the Fourier Transform Method and the LPN Method for a linear 50 Ω load. These calculations were performed for the antenna parameters, angle of arrival and incident field shown in the figure.

EMP Excitation of a Monopole Antenna

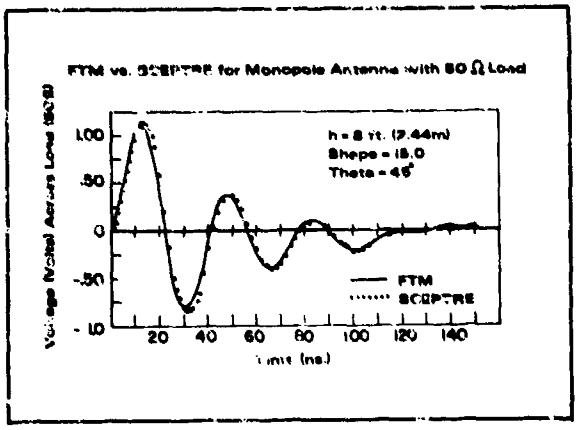


An LPN representation of this antenna is shown. The element values of the circuit model have been obtained from available tables.

Monopole Antenna Equivalent Circuit

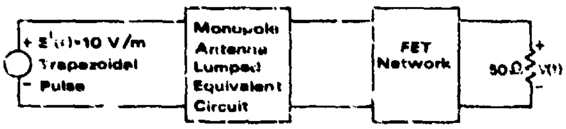


The results of the analysis using SCEPTRE are compared to those obtained from the Fourier Transform Method.

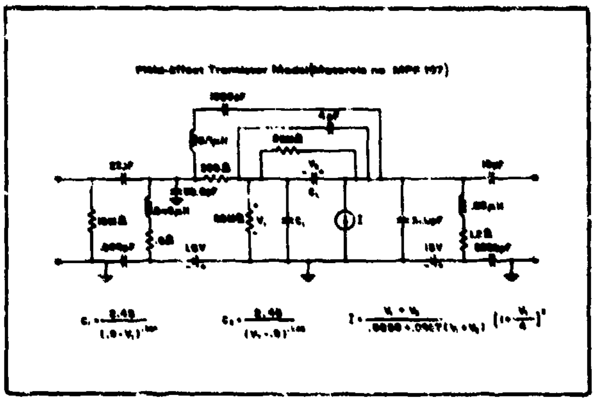


A second example deals with the analysis for a nonlinear load. In this example, an 8 foot monopole over a perfectly conducting ground is loaded with a field-effect transistor (FET) amplifier. The input trapezoidal pulse is limited to a 10 V/m maximum amplitude to keep the gate bias reversed, since the FET network model is not valid when the gate is forward biased. The purpose of this example is to demonstrate the implementation of SCEPTRE using a receiving monopole connected to a receiver front end.

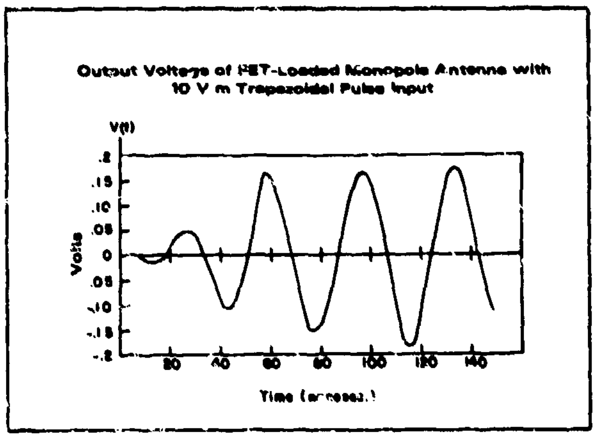
FET-Loaded Monopole Antenna



This model of the FET is used in the calculations.



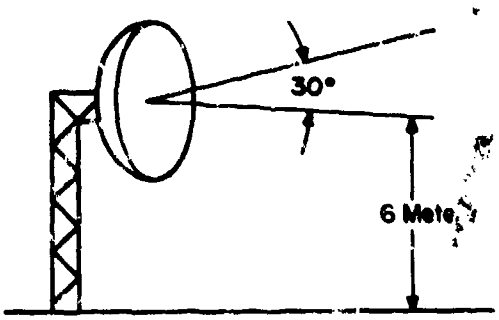
The response of the FET amplifier is shown here.



This shows the early time response. The ring up is due to the Q of the circuit. A similar ring down would be seen in late time.

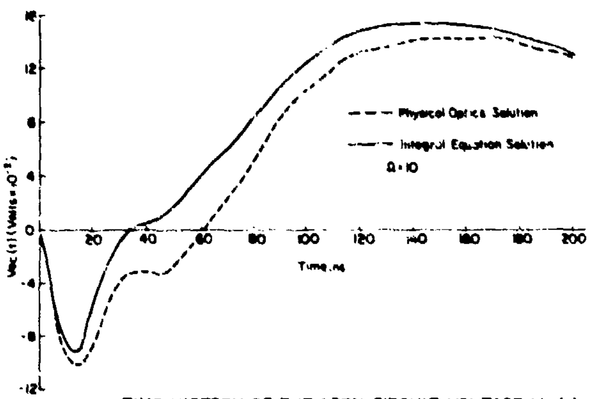
Parabolic Antenna Example

The EMP response of a parabolic reflector antenna with a dipole feed has been considered. The basic geometry considered was a 25 foot paraboloidal antenna, 6 meters above ground with its axis tilted 30 degrees with respect to the ground.

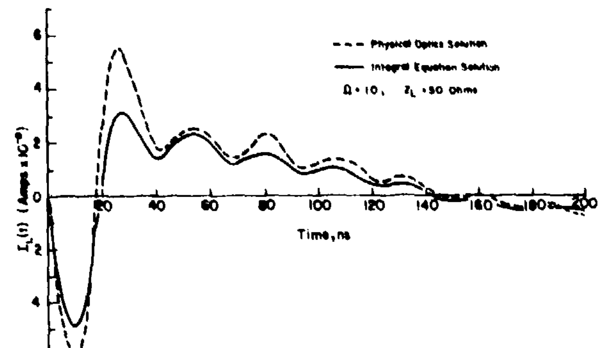


Integral equations for the current distribution on a parabolic cylindrical reflector were derived using superposition methods. These integral equations were solved numerically using an iterative method. As expected, the frequency domain focal-plane scattered fields are poorly focused for most frequencies of the EMP spectrum. This implies that the feed antenna sees a practically uniform incident field.

The calculated open circuit voltage and load current for this case are shown in the following figures for both the over ground and free space conditions.



TIME HISTORY OF THE OPEN CIRCUIT VOLTAGE, $V_{oc}(t)$, FOR A RECEIVING DIPOLE FEED ANTENNA TO HIGH ALTITUDE EMP



TIME HISTORY OF THE LOAD CURRENT, $I_L(t)$, FOR A RECEIVING DIPOLE FEED ANTENNA IN FREE SPACE TO HIGH ALTITUDE EMP

4.6 STRUCTURES MODELED AS ANTENNAS

Missile in Flight

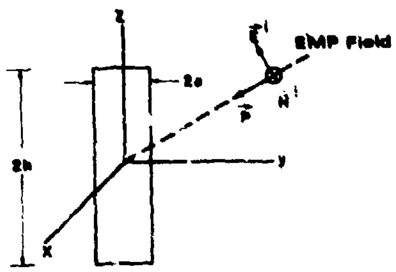
A useful model for a missile in flight is a short circuited cylindrical dipole antenna. For shape factors (Ω) greater than ten (10), the missile can be considered as a thin dipole antenna and solved by the Fourier Transform techniques.

The current distribution induced by an incident EMP on the surface of a missile body can, in principle, be determined on the basis of scattering or antenna theory. The computational effort required depends to a large degree on the analytical model and degree of accuracy required. In general, the current density on the surface of a missile structure has two tangential components (for a perfectly conducting surface) which are related through a system of two coupled integral equations. Solutions to such integral equations are indeed very difficult and will not be considered here. A great deal of simplification results if a cylindrical model for the missile structure is chosen. It is usually assumed that no circumferential currents are induced, which is valid provided the diameter of the cylindrical antenna is less than a quarter wavelength. It should be recognized that for structures whose length-to-diameter ratio is not large (shape factor > 10), the circumferential currents can be as large as the axial currents. Moreover, first order approximations to the current (axial) distribution are possible using thin linear antenna theory. This antenna theory requires that

$$h \gg a \text{ and } 2a/\lambda \ll 1$$

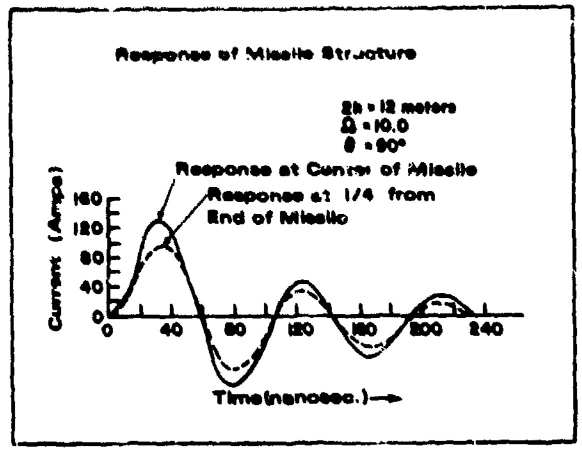
If these conditions are violated, the resulting current distribution is expected to be only a rough estimate of the actual one. Fair agreement (a factor of 2) in terms of peak current can sometimes be obtained. The major discrepancy occurs in the actual distribution of the currents on a complex structure due to the superposition of the various components comprising the total current.

Model for Missile Structure



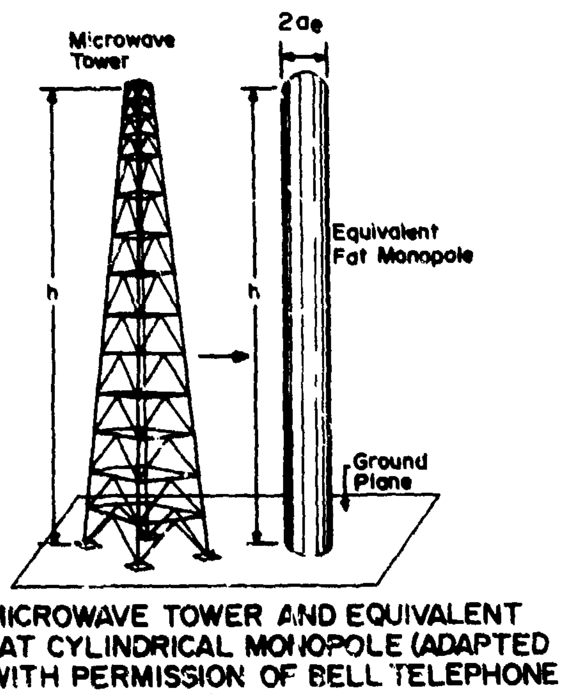
The predicted response of a 12-meter cylindrical structure to a 10,000 V/m double-exponential environment for broadside incidence are presented in the figure. The shape factor of the structure is Ω = 10.0 which corresponds to a length-to-diameter ratio of 75.0. As expected, the amplitude of the current at the center of the antenna is greater than that three meters from the center. Otherwise, the current waveforms are almost identical. Note that the fundamental frequency of the current ringing is 11.2 MHz whose period is 89 μsec and is approximately the time it takes a current wave to traverse the length of the antenna twice.

Having determined the missile skin current, the next step is to calculate the electric field on the inside surface of the missile skin through its surface transfer impedance, as discussed later. This surface field may be viewed as a distributed source, which excites the interior of the missile body, resulting in voltages and currents appearing at various terminals of electronic equipment. The missile skin current also contributes to the antenna excitation and can couple to the interior via apertures or penetrants in the missile skin.



Microwave Tower

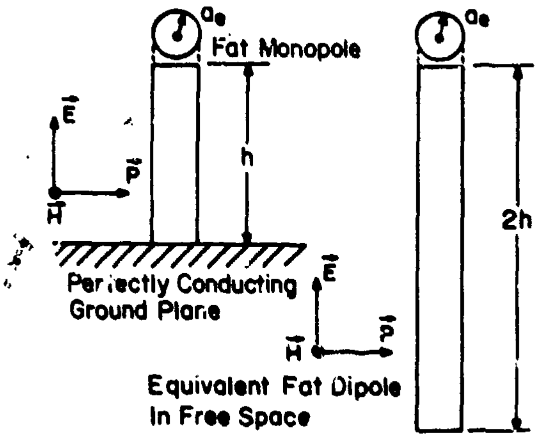
An open type microwave tower structure can be approximately modeled as a fat monopole to calculate the tower current due to EMP. An effective radius (a_e) can be determined for the monopole which depends on the tower geometry. Assuming a cylinder (and, consequently, a_e) that is the same size as the base of the tower gives an upper bound on the tower current. A lower bound is obtained by assuming a cylinder equivalent to the size of the waveguide. The difference between the upper and lower bounds is approximately a factor of 3.



MICROWAVE TOWER AND EQUIVALENT FAT CYLINDRICAL MONOPOLE (ADAPTED WITH PERMISSION OF BELL TELEPHONE LABORATORIES)

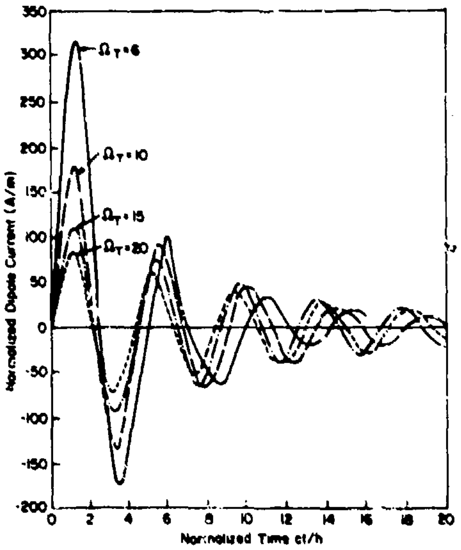
To simplify the calculation and determine an upper bound for the tower current two assumptions are made: (1) tower monopole and earth (ground plane) are perfectly conducting; and (2) that the incident EMP electric field vector is parallel to the axis of the cylinder. The effect of these assumptions is to provide an upper bound which is within a factor of approximately 2.

For purposes of analysis, a monopole which is perfectly conducting over a perfectly conducting ground plane can be considered as a dipole of half-height equal to the height of the monopole.



TOWER MONOPOLE AND EQUIVALENT FAT DIPOLE (ADAPTED WITH PERMISSION OF BELL TELEPHONE LABS)

The approximate EMP induced current normalized to the dipole half-height on a normalized time scale for four shape factors is shown.



EMP-INDUCED TOWER CURRENT (ADAPTED WITH PERMISSION OF BELL TELEPHONE LABORATORIES)

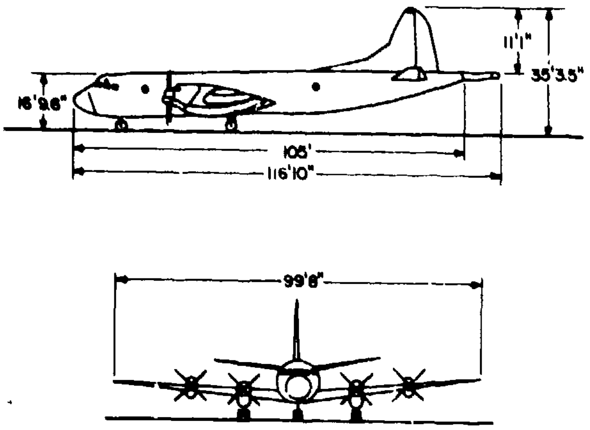
The ringing frequency is approximately equal to the fundamental resonant frequency of the dipole.

P-3C Aircraft

Another example of EMP predictive modeling has been the P-3C Aircraft. Typical modeling of aircraft utilizes thin wires. The simplest approximation is the wire cross model where the fuselage and wings are modeled by wires, neglecting the tail structure.

This was done for a P-3C aircraft as depicted below.

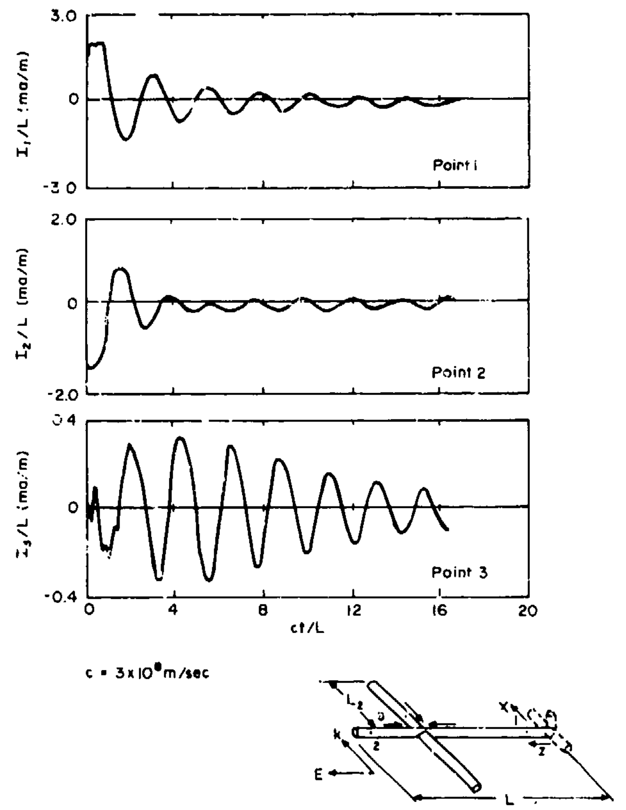
The shape factor used for this analysis was $\Omega = 2 \ln L/a = 7.0$ (a = wire radius). The wire radius used was somewhat smaller than the average occurring on the real aircraft due to numerical calculation difficulties. This results in slightly higher peak currents due to the higher Q associated with the smaller radius.



P-3C Aircraft Overall Dimensions

The wire model and calculated response data are shown in the following figure. The data points are located at: (1) Point 1 at $Z/L = 0.47$, (2) Point 2 at $Z/L = 0.71$, and (3) Point 3 at $X/L_2 = 0.55$. The excitation used was a 50 kV/m step drive field.

The analysis was performed using dipole models in free space for each wire of the cross wire model. Calculations of this type have shown reasonable agreement in terms of principal ring frequencies and to a lesser, but acceptable, accuracy for peak currents and current distributions with experimental data. The circumferential currents are not predicted using the thin wire model approximation.



Wire cross currents at points 1-3 (fuselage excited by a unit step electric field.)

4.7 CABLE ANALYSIS

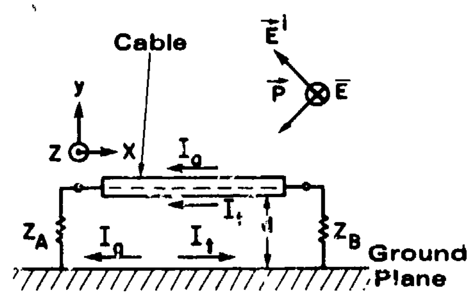
As in the case of antenna coupling, to determine the load voltage, current, or energy due to EMP excitation of cables, it is necessary to calculate the current voltage distribution on the cable. Depending on the cable construction and physical configuration, the current distribution can be determined by either antenna theory or transmission line theory.

In this section, we will look at simple cables and geometries to illustrate the analytical methods employed.

Cables in Proximity to Conducting Surfaces

In general, for cables in close proximity ($d << \lambda$) to a conducting surface (i.e., earth or other conducting surface) antenna currents can be neglected. Due to the reflection from the conducting surface in close proximity, the net field at the cable for use in antenna theory is nearly zero. The cable current (for unshielded cables) or the sheath current (for shielded cables) is totally due to transmission line currents.

Nature of Current Distribution on Cables



Total Sheath Currents

$I(X) = I_a + I_t$

I_a = Antenna Current

I_t = Transmission Line Current

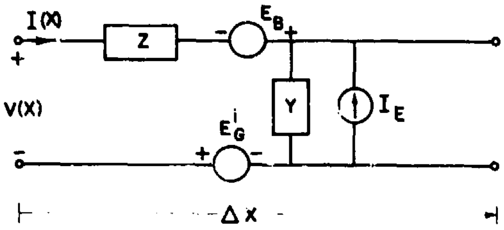
$I_t \approx 0$ for $d >> \lambda$

$I_a \approx 0$ for $d << \lambda$

Horizontal Cable Over Ground

The transmission line equivalent circuit of a small cable section over a finitely conducting ground plane is shown in the figure.

Incremental Equivalent Circuit of a Horizontal Cable Configuration



$$I_E = Y\int_0^d E_y^i\, dy$$

$$E_B = j\omega\mu\int_0^s H_z^i\, dy$$

E_G^i = Tangential Electric Field at the Surface of the Earth with No Cable Present

The transmission line equations are given by:

$$-\frac{\partial V(x)}{\partial x} = ZI(x) - E_B - E_G^i$$

$$-\frac{\partial I(x)}{\partial x} = YV(x) - I_E$$

where

Z = cable impedance per unit length in proximity to the ground plane

Y = cable admittance per unit length in proximity to the ground plane

E_G^i = tangential electric field at the surface of the ground plane and in the absence of the cable

$V(x)$ = voltage at point x due to incremental sources

$I(x)$ = current at point x due to incremental sources

E_B = incremental voltage source

I_E = incremental current source

It is to be noted that in this formulation antenna currents have been neglected.

If the incident EMP field is known, the point-source generators shown can be evaluated by

$$I_E(x) = Y\int_0^d E_y^i\,(x,y)\,dy$$

$$E_B(x) = j\omega\mu\int_0^d H_z^i\,(x,y)\,dy$$

The transmission line equations are solved for a point voltage and current generator located at some general point $x = \xi$ along the cable. The result will give the Green's function solution, $G(x, \xi)$, to this problem for prescribed loading conditions (Z_A, Z_B). Then the total current, $I(x)$, at any point along the cable is obtainable by use of the superposition integral:

$$I(x) = \int_0^L G_I(x, \xi)\, I_E(\xi)\, d\xi + \int_0^L G_V(x,\xi)\left[E_B(\xi) + E_G(\xi)\right] d\xi$$

where

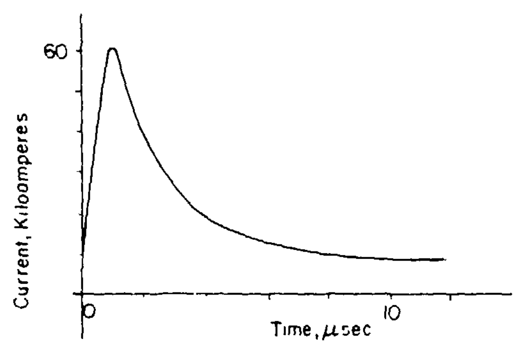
$G_I(x,\xi)$ = Green's function due to a point current source

$G_V(x,\xi)$ = Green's function due to a point voltage source

L = length of cable.

The Green's functions are the solution of the transmission line equations for a point voltage source or current source at a point "ξ" on the line. The total current on the line is then the summation (integral) of the contributions of each of the point sources.

A typical time history of the current in a long cable above ground is shown in the figure. This waveform was produced for a vertical electric field at a grazing angle of incidence. The time history exhibits a high-amplitude spike with a long decay (tail). This indicates the low-frequency pickup is highly important.

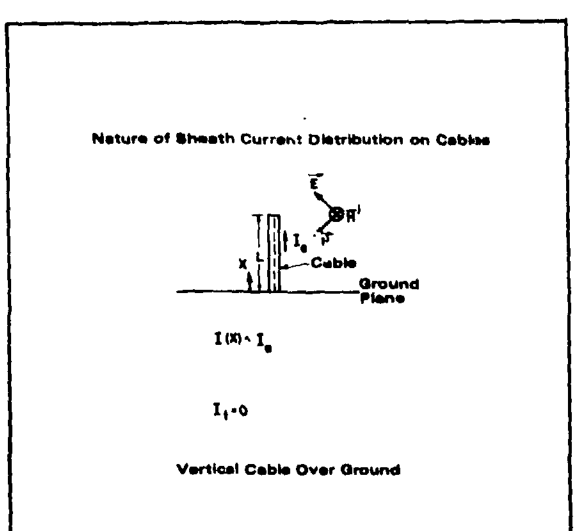


WIRE ABOVE GROUND · VERTICAL E FIELD

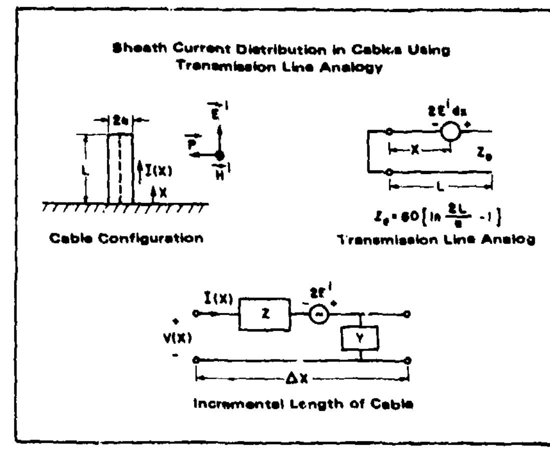
Cables in Proximity to a Non-Conducting Surface

For cable runs which are close to non-conducting surfaces, or far removed from conducting surfaces, the transmission line currents are quite small (negligible) compared to the antenna currents.

To determine the sheath current for shielded cables, or the pickup by unshielded cables (single wires), the cable can be viewed as an equivalent dipole. In this case, the antenna analysis discussed previously, can be used to determine the current distribution.



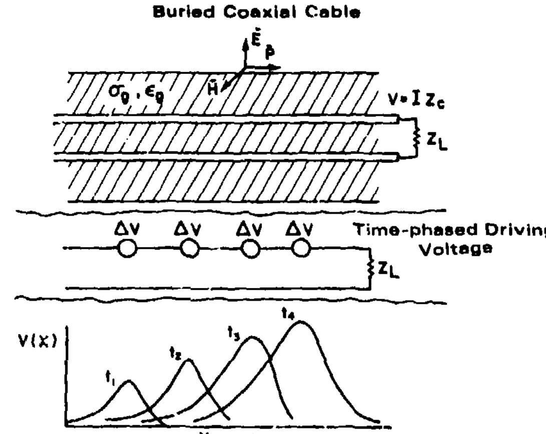
If the antenna current distribution on a cable configuration is not available from antenna theory, the transmission line approach may be used to calculate the approximate current distribution. This is based on the close analogy that exists between wire antennas and transmission lines. This analogy is shown in the figure.



Buried Cables

For buried cables, the principal pickup mechanism to cause sheath currents is the common impedance mechanism. Electric and magnetic fields cause currents to flow in the earth, and the resistance of the earth causes a voltage drop to appear along the cable.

Buried Coaxial Cable



For buried cables, the current induced is obtained from a transmission line model in which the soil surrounding the cable is the return conductor. To model this configuration as a transmission line, it is necessary to have the characteristic impedance (Z_o) and propagation (γ) of the cable. These are given by:

$$\gamma = \sqrt{ZY}$$

$$Z_o = \sqrt{Z/Y}$$

$$Z = Z_g + Z_i + j\omega L$$

$$Y = j\omega C\ Y_g / j\omega C + Y_g$$

where

Zg = soil impedance

Z_i = cable impedance

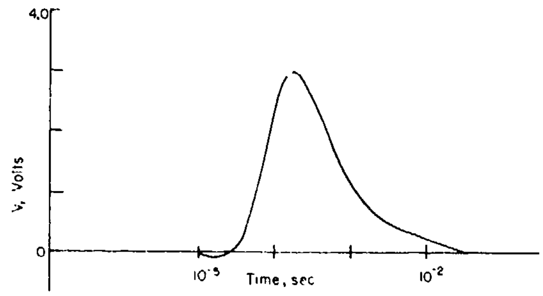
$j\omega L$ = inductive reactance of insulation gap

$j\omega C$ = capacitive reactance of insulation

Y_g = soil admittance.

For near surface burial (a few meters) the incident field is approximately that at the earth's surface. For deep burial, the propagation loss in the earth must be taken into account. The surface field or transmitted field can be determined from the reflection and transmission coefficients, given in a later section, and the earth propagation constant.

Depending on the angle of arrival, the incremental voltage drops appear to be progressively excited along the cables so that a time-phased effect of the driving voltage must be considered. In the case of a tangential angle of arrival with the direction of propagation parallel to the axis of the buried cable, there is a tendency toward a traveling wave buildup of the sheath current or the cable as the wave progresses along the cable. Fortunately, in general, this is counteracted by the differences in propagation time in the earth and in the wave above the earth. The peak amplitude and energy are greatly reduced by earth absorption. Due to the earth parameters, the time history of the current in the cable (deep burial > 10 meters), indicates the primary pickup is low frequency and of low amplitude.



WIRE BELOW GROUND - HORIZONTAL E FIELD

Surface Transfer Impedance for Shielded Cables

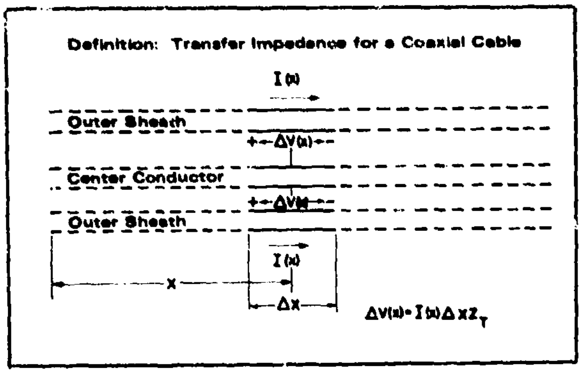
Once the sheath current distribution is known, the voltage induced into a shielded cable because of imperfect shielding can be calculated.

The shielding effectiveness of a cable can be represented quantitatively by surface transfer impedance, which relates the sheath current flowing on the cable shield to the voltage drop per unit length (surface electric field) appearing at the inner surface of the shield.

A current, I_S, flowing on the outer sheath of the cable causes an incremental voltage drop, ΔV, to appear across an incremental length, Δx, on the inside of the sheath. This voltage is given by

$$\Delta V = Z_T I_S(x)\Delta x$$

where Z_T is defined as the surface transfer impedance of the cable. The surface transfer impedance is determined by the construction of the outer shield. Analytical expressions are available for solid-shell and braided coaxial cables. Since braided cables present a geometry that is quite difficult to analyze in detail, their transfer impedance is most easily determined experimentally.



Before presenting expressions for the surface transfer impedance of a solid-state coaxial cable, it is useful to consider the problem qualitatively. In the limit of zero frequency, a current flowing on the outer shell of a coaxial cable will see the dc resistance of the shell. At very low frequencies and thin wall shields, there is little attenuation due to skin effect. Therefore, at zero frequency, the voltage drop appearing inside the shell will be the shell current multiplied by the dc resistance of the shell given by

$$R_{dc} = \frac{1}{2\pi b t \sigma} \text{ ohms/meter}$$

for thin-wall shells, where

- σ = the conductivity of the outer conductor
- t = the thickness of the outer conductor
- b = the inside radius of the outer conductor.

As the frequency of the shell current is raised, less and less of it will penetrate the shell, and thus one would expect the surface transfer impedance of the solid-shell coaxial cable to become smaller.

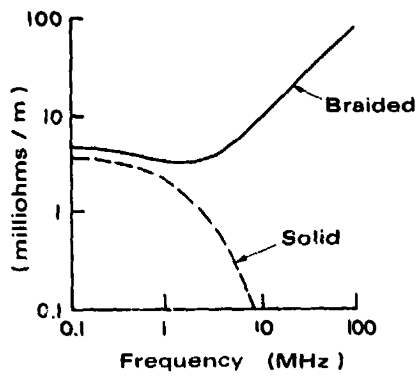
Schelkunoff has developed an approximate expression for the surface transfer impedance of a solid-shell coaxial cable using the assumption that the thickness of the shell, t, is much less than its inner radius. It is of the form

$$Z_t = \frac{\eta}{2\pi \sqrt{bc}\ \sin h\ (\Gamma t)} \quad \text{for } t \ll b$$

where

- η = intrinsic impedance of the shield
- Γ = propagation constant of the medium
- c = outer radius of outer conductor
- b = inner radius of outer conductor.

Surface Transfer Impedance of Braided and Solid Outer Conductor Coaxial Cable



The figure also shows the surface transfer impedance for a typical coaxial cable with a braided outer conductor. Note that at low frequencies, i.e., less than 1 MHz, the behavior tends to follow that of the solid outer conductor, and at zero frequency the surface transfer impedance is given by the dc resistance per unit length of the wires that make up the braid. Above about 2 or 3 Mhz, the surface transfer impedance begins increasing with frequency. Kruegel investigated braided coaxial lines in great depth. He found that this high-frequency behavior was strongly influenced by details of the braid construction -- optical covering factor, number of carriers, and braid angle to name three important factors. These factors determine the size and shape of the holes in the braid and, thus, the magnitude of the magnetic fields which can fringe into the interior of the cable. It is apparently these fringing fields which determine the high-frequency behavior of the cable.

Based on experimental results, the surface transfer impedance for braided shell coaxial cables is characterized by a diffusion term representing diffusion of EM energy through the metal and an inductance term representing penetration of the magnetic field through the holes.

$$Z_T = Z_D + j\omega M_{12}$$

where

Z_D = diffusion term equal to the dc resistance of the braid per unit length at low frequencies

M_{12} = the leakage mutual inductance of the braid per unit length (may be positive or negative).

The diffusion term, Z_D is given by

$$Z_D \approx \frac{4}{\pi d^2 \; N \; C \; \sigma \; \cos\alpha} + \frac{(1+j)\; d/\delta}{\sinh\;(1+j)\; d/\delta}$$

and the mutual inductance term is given by:

$$M_{12} \approx \frac{\pi \mu_o}{6C} (1 - K)^{3/2} + \frac{e^2}{E(e) - (1-e^2)\; K\;(e)} \qquad \alpha < 45^o$$

$$\approx \frac{\pi \mu_o}{6C} (1 - K)^{3/2} + \frac{e^2 \quad \sqrt{1-e^2}}{K(e) - E(e)} \qquad \alpha > 45^o$$

where

K = optical coverage

C = number of carriers

N = number of ends

d = diameter of individual wires

α = weave angle

δ = skin depth = $\frac{1}{\sqrt{\pi f \mu \sigma}}$

σ = wire conductivity

μ_o = free space permeability = $4\pi \times 10^{-7}$

K(e) = complete elliptic integral of the first kind

E(e) = complete elliptic integral of the second kind

$$e = \sqrt{1-\tan^2\alpha} \qquad \alpha < 45^o$$

$$= \sqrt{1-\cot^2\alpha} \qquad \alpha > 45^o$$

These expressions have been found to be very accurate at low frequencies ($d/\delta << 1$) and accurate to within a factor of 3 or less at high frequencies ($\omega M_{12} >> |Z_D|$).

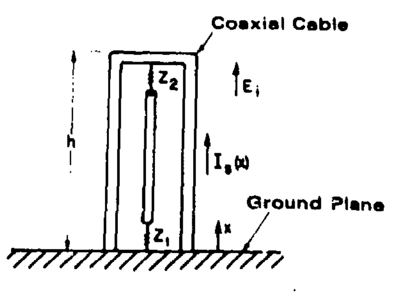
Several coaxial cable types have been investigated to determine their surface transfer impedance. The results for two of the more common cable types are given below:

Cable Type	Resistance (ohms/m)	Inductance (Hz/m)
RG-8A/U	4.5×10^{-3}	$+ 8.75 \times 10^{-1}$
RG-9A/U	3.2×10^{-3}	$- 1.91 \times 10^{-1}$

EMP Response of Typical Cable

We will now consider an example of EMP penetration into a coaxial cable. Let a coaxial cable be exposed to an incident field, E_i, parallel to its axis. With the outer conductor considered as a linear antenna, the external incident field induces an axial current distribution on this conductor. If the current, $I_s(x)$, is known as a function of x, the coupling into the cable can be represented by a continuous distribution of incremental generators each with voltage $Z_T I_S(x)dx$, where Z_T is the surface transfer impedance of the cable shield. To determine the current distribution inside the cable from which the load current through Z_1 can be obtained, it is necessary to have expressions for the current and voltage at any point on the line with arbitrary loads (Z_1, Z_2) for an arbitrary location of a series point generator; this is the Green's function solution to the problem. Then, by the superposition integral, the current distribution due to a voltage source distribution is readily obtained.

Loaded Coaxial Cable in an Incident Electric Field



Let a section of the line h be terminated in Z_1 and Z_2, and $V_1(x,\xi)$ and $I_1(x, \xi)$ be the voltage and current at point x when a point source, $V_g = Z_T I_S(\xi)\, d\xi$, is impressed at $x = \xi$ in series with the line. Note that this transmission line corresponds to the interior of the coaxial cable. The voltages and currents in this line are given by Schelkunoff. Let the current Green's function be expressed as:

$$G_I(x,\xi) = I_1(x,\xi)$$

Then by the superposition integral

$$I(x) = \int_0^h G_I(x,\xi)\, I_s(\xi)\, d\xi$$

and

$$I(o) = I_L =$$

$$\int_0^h G_I(o,\xi)\, I_s(\xi)\, d\xi .$$

In order to perform this integration, the sheath current distribution, $I_s(\xi)$, must be known. Since the shield (outer conductor) of common coaxial cables satisfies the conditions of linear antenna theory ($h/a >> 1$, $2a/\lambda << 1$), it may be treated as an unloaded receiving antenna in a uniform field. The total axial current (sheath current) distribution is approximated by the following expression:

$$I_s(\xi) = I_s(o) \frac{\cos\beta\xi - \cos\beta h}{1 - \cos\beta h}$$

where

β = propagation constant of the medium surrounding the antenna (cable)

$$I_s(o) = -h_e E_i(j\omega)/Za.$$

In this relationship, h_e is the effective length of a monopole antenna of physical length, h, and Z_a is its input impedance. In the equation concerning the frequency-domain transfer function, the current in the load, Z_1, is related to the incident electric field $E^i(j\omega)$, the time history of the load current is determined by taking the inverse Fourier transform. Let

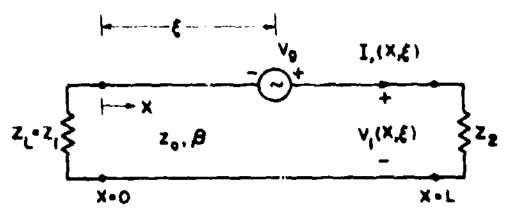
$$I_L(\omega) = I_{LR}(\omega) + j\, I_{LI}(\omega)$$

be the spectrum of the load current decomposed into its real and imaginary parts. The time history of the load current is then

$$i_L(t) = \frac{1}{\pi}\int_0^{\omega_c} [I_{LR}(\omega)\cos\omega t - I_{LI}(\omega)\sin\omega t]\,d\omega$$

where, in obtaining the second part, use has been made of the relation $I_L(j\omega) = I_L(-j\omega)$, which is the case for any time invariant, linear system. The radian cut-off frequency, ω_c, is used for computational purposes and is determined on the basis of the high-frequency content of the excitation pulse. For example, the highest significant frequency contained in a Gaussian pulse is usually taken to be $f_c = 2.6\, f_1$, where $f_1 = 1/2\pi t_1$ and t_1 is a measure of the pulse width, ($t_1 = 0.4246\, t_\omega$ where t_ω is the pulse width at the half amplitude points). Numerical techniques may be used to integrate this equation with the aid of a digital computer.

Transmission Line with a Voltage Point Source



The description of the incident electric field pulse assumed here is a Gaussian pulse of unit amplitude and of width t_ω = 100 nsec at 1/2 amplitude point. It is given by the expression

$$e_i'(t) = \exp\left[-\frac{t^2}{2t_1^2}\right]$$

where $t_1 = 0.4246\, t\omega$ sec is a measure of the pulse width. The spectrum of the pulse described is

$$E_i(j\omega) = \frac{\sqrt{2\pi}}{\omega_1}\exp\left[-\frac{\omega^2}{2\omega_1^2}\right]$$

where $\omega_1 = 2\pi f_1 = 1/t_1$ rad/sec. For this example, a 25-meter RG-8A/U and a 25-meter RG-9A/U cable were considered. The electric field vector was parallel to the axis of the cables and broadside incident. The pulse width assumed corresponds to a cutoff frequency $f_c = 9.75$ MHz.

The transient response of the load current was obtained numerically for the following special cases:

Case I

RG-8A/U Cable

$h = 25$ meters

$t_\omega = 100$ nsec

$Z_1 = Z_o = 50$ ohms

$Z_2 = 0$

Case II

RG-9A/U Cable

$h = 25$ meters

$t_\omega = 100$ nsec

$Z_1 = Z_o = 50$ ohms

$Z_2 = 0$

Case III

RG-8A/U Cable

$h = 25$ meters

$t_\omega = 100$ nsec

$Z_1 = Z_o = 50$ ohms

$Z_2 = \infty$

Case IV

RG-9A/U Cable

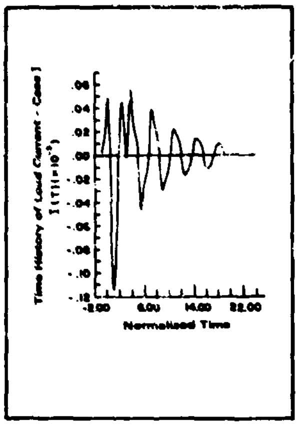
$h = 25$ meters

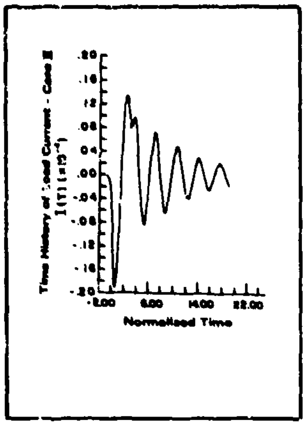
$t_\omega = 100$ nsec

$Z_1 = Z_o = 50$ ohms

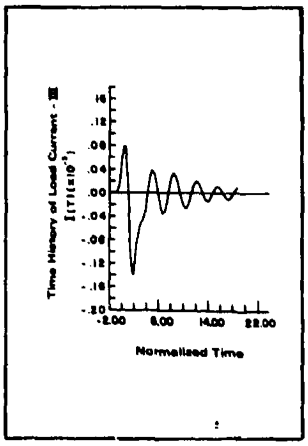
$Z_2 = \infty$

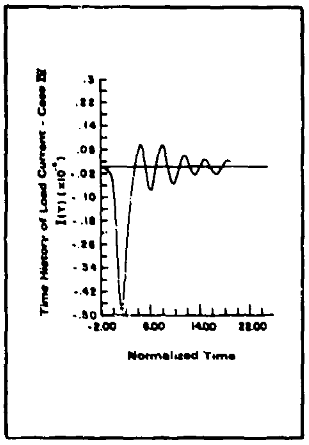
Shown are the time histories of the load currents for Cases I and II for a Gaussian pulse whose amplitude is 1V/m defined previously. Note that the time scale is normalized with respect to the pulse width. The results indicate that the ringing of the transient response is mainly due to the fundamental resonant frequency of the cable when viewed externally as an unloaded scattering antenna. The fundamental period of current oscillations is determined by the time it takes the current wave to travel four times the cable length, with the propagation velocity of free space.





The time histories for Cases III and IV are shown here.





4.8 SHIELDING ANALYSIS

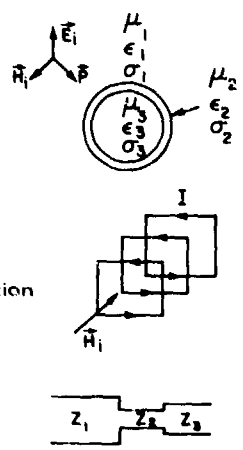
There are about as many analytical approaches to calculating shielding effectiveness as there are shielding engineers. In general, the rigorous approaches involve some simplifying assumptions which are sometimes pertinent to the EMP problem area. In this case, they assume the conductivity of the shield to be such as to permit first solving for the current distribution on the exterior of the shield, as we have done in the previous sections. It also assumes that the shield is reasonably good so that any equipment configurations inside the shield are not a dominant factor.

Thus, the shielding approaches have been boiled down to the three basic groups; exact approach based on scattering theory, approaches which involve some implicit assumptions which can be shown to give rise to the lumped circuit approximation, and the Schelkunoff plane wave approach which gives rise to the so-called transmission line equivalent. The rigorous or lumped-circuit approximation appears to be a satisfactory approach for most EMP-type shielding applications. It does tend to break down where the conductivity of the wall material is low; for example, the conductivity of coke or wall material made out of seawater.

Scattering theory

Lumped-circuit approximation

Plane wave, transmission line equivalent circuit



Scattering Theory Solutions

For simple geometries such as spherical, and cylindrical shells and parallel plates, scattering theory solutions have been obtained. These are presented in the form of transfer functions as a function of frequency.

Based on the exact scattering theory, for frequencies greater than a few Hertz, the transfer function for magnetic field shielding is of the form:

$$T_H(\omega) \equiv \frac{H_{in}(\omega)}{H_{ex}(\omega)}$$

and is given by:

$$T_H(\omega) = \frac{1}{\cos(k_2 d) - \frac{k_2 b}{2}\sin(k_2 d)}$$

ω = $2\pi f$

f = frequency of incident field

k = $(\mu_o/\mu)k_2 b$

k_2 = $\sqrt{-j\omega\mu\sigma}$ = $\frac{\sqrt{-j2}}{\delta}$

δ = $1\sqrt{\pi f\mu\sigma}$ = skin depth of material

d = thickness of enclosure walls

μ_o = permeability of free space

μ = permeability of enclosure walls

σ = conductivity of enclosure walls

The factor "b" in the equation is a geometric variable that characterizes different enclosure geometries as follows:

b = the separation distance between plates for large area parallel plate shields

b ' = the radius for cylindrical enclosures

b = 2/3 the radius for spherical enclosures

At high frequencies, where the wall thickness is greater than the skin depth ($d > \delta$), the transfer function reduces to

$$T_H(\omega) = \frac{2\sqrt{2}\,\delta\, e^{-d/\delta}}{b}$$

At low frequencies ($d < \delta$) the magnetic shielding becomes

$$T_H(\omega) = \left| \frac{1}{\frac{j\omega d\sigma\mu b}{2} + 1} \right|$$

The transfer function for electric field shielding is given by:

$$T_E(\omega) = \frac{2\,(k_1 b)^2}{k \sin k_2 d}$$

$k_1 = 2\pi/\lambda$

λ = wavelength of impinging field

for $d > \delta$, high frequencies:

$$T_E(\omega) = \frac{9\,\omega\varepsilon_o b e^{-d/\delta}}{\sqrt{2}\,\sigma\delta}$$

$\varepsilon_o = 10^{-9}/36\pi$ farads/meter

for low frequencies, $d < \delta$:

$$T_E(\omega) = \frac{9\omega\varepsilon_o b}{4\sigma d}$$

A conservative approximation for other geometries (cubes, rectangles, etc.), modeling a structure as a sphere with b being the minimum dimension of the enclosure is good practice.

Low Frequency Lumped Circuit Approximation

The very low frequency magnetic field penetration characterization for a sphere is given in the form of an R-L circuit. The shield is regarded as a good antenna with inductance L. The Thevenin equivalent voltage generator is equal to the magnetic field intensity incident on the shield. The various circuit parameters are related to the parameters of the spherical shell:

$$R = \frac{2\pi n^2}{3 d_\tau}$$

$$L = \frac{2\pi\mu a n^2}{9}$$

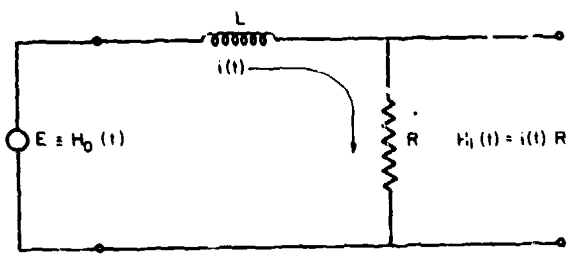
a = radius of sphere

d = wall thickness

σ = wall conductivity

μ = free space permeability

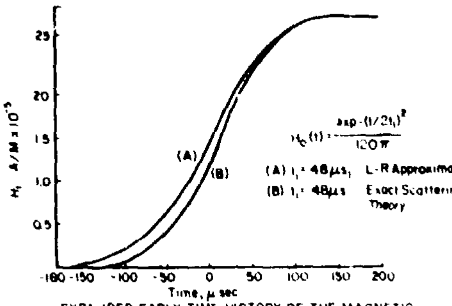
n = equivalent number of turns (this cancels out in the final expression for shielding effectiveness).



ANALOG CIRCUIT CHARACTERIZING SHIELDING EFFECTIVENESS OF SPHERES

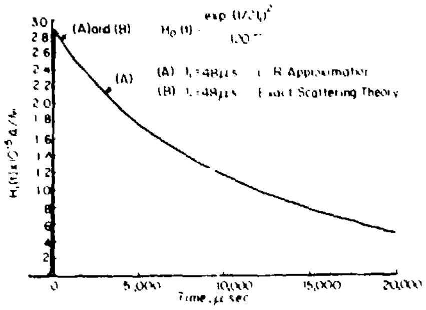
Using the circuit approximation, the time history of interior field with a Gaussian pulse of $120/\pi$, amperes/meter peak amplitude incident on the shield was calculated. The shield was an 18 inch radius, 1/16 inch wall thickness aluminum sphere. The calculation, using the circuit approximation, was compared to one using exact scattering theory.

The early time-history of the interior field is shown in the figure. The maximum interior magnetic field intensity is proportional to the integral of the incident field. This maximum interior field is reached with a rise time approximately equal to the incident field duration.



EXPANDED EARLY TIME HISTORY OF THE MAGNETIC FIELD INTENSITY AT THE CENTER OF AN 18 inch RADIUS ALUMINUM SPHERE WITH A 1/16 inch WALL TO A GAUSSIAN PULSE

An exponential decay describes the long time response after the initial rise. The decay time constant is R/L.



LONG TERM HISTORY OF THE MAGNETIC FIELD INTENSITY AT THE CENTER OF AN 18 inch ALUMINUM SPHERE WITH A 1/16 inch WALL

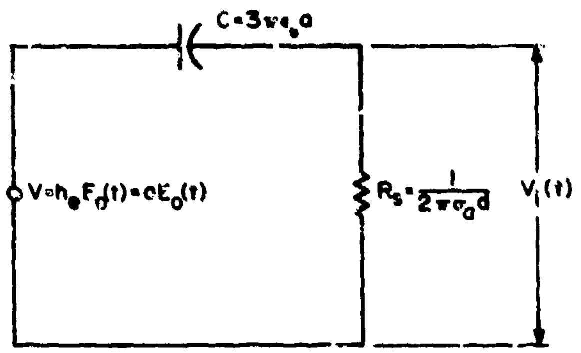
The very low frequency electric field penetration characterization for a sphere is given in the form of an R-C circuit. The shield is regarded as a dipole antenna with effective height h_e and capacitance C. The Thevenin equivalent voltage generator is proportional to the electric field intensity incident on the shield. The various circuit parameters are related to the parameters of the spherical shield:

a = radius

d = wall thickness

σ_e = wall conductivity

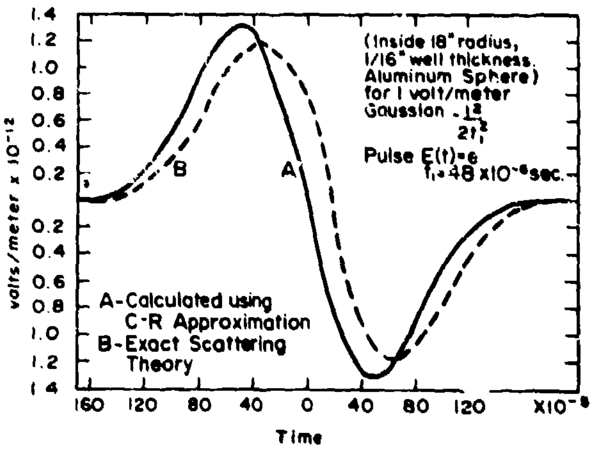
The electric field at the center of the sphere is proportional to the voltage across R_s.



For $a >> d$ and $\frac{2\sigma_0 d}{3\epsilon_0} > \omega$

VERY LOW FREQUENCY ELECTRIC FIELD PENETRATION CHARACTERIZATION FOR A SPHERE

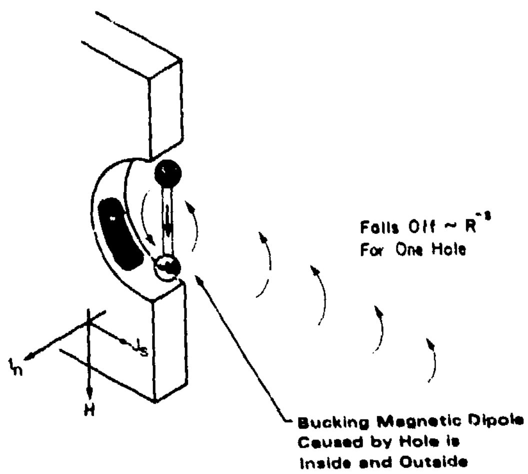
Using the low frequency circuit approximation, the time history of the interior field with a Gaussian pulse of 1 volt/meter peak amplitude incident on the shield was calculated. The shield was an 18 inch radius, 1/16 inch wall thickness aluminum sphere. The calculation using the circuit approximation was compared to one using exact scattering theory.



TIME HISTORY OF THE ELECTRIC-FIELD INTENSITY AT THE CENTER OF A SPHERE

Apertures

Obviously, if the enclosure has holes, this provides a means for the exterior fields to leak into the interior. Typically, if we assume the enclosure aperture to be very small compared to the general size of the enclosure, the effect of a small aperture can be calculated. In this case, a magnetic dipole is assumed to appear in the aperture hole which is polarized in such a way as to cancel the exterior current flow. However, since this dipole is coupled both to the exterior and interior, this provides a source of interior fields which tends to fall off at the low frequencies inversely proportional to the cube of the distance away from the aperture. In the case of real enclosures, the effect of the enclosure walls must also be considered.



For a plane wave incident on the structure, both the electric and magnetic fields will penetrate. The EMP propagation is in a direction parallel to the shielding plane. The geometry is depicted in the figure. The penetrating fields are given by:

$$E_r = 2/3\pi \left(\frac{a}{r}\right)^3 E_o \cos\theta$$

$$E_\theta = \frac{1}{3\pi} \left(\frac{a}{r}\right)^3 E_o \sin\theta$$

$$E_\phi = 0$$

$$H_r = \frac{4}{3\pi} \left(\frac{a}{r}\right)^3 H_o \sin\phi \sin\theta$$

$$H_\phi = \frac{2}{3\pi} \left(\frac{a}{r}\right)^3 H_o \cos\phi$$

$$H_\theta = \frac{2}{3\pi} \left(\frac{a}{r}\right)^3 H_o \sin\phi \cos\theta$$

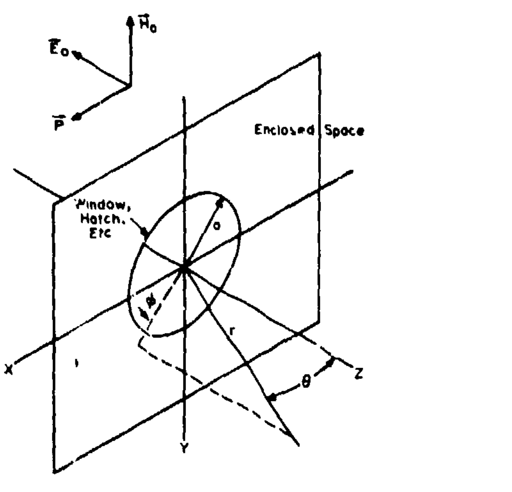
where

E_ϕ, E_r, E_θ = internal electric field components in spherical coordinates

H_ϕ, H_r, H_θ = internal magnetic field components in spherical coordinates.

a = radius of the aperture. In analysis of non-circular apertures, "a" should be set equal to 1/2 the largest dimension of the opening under consideration

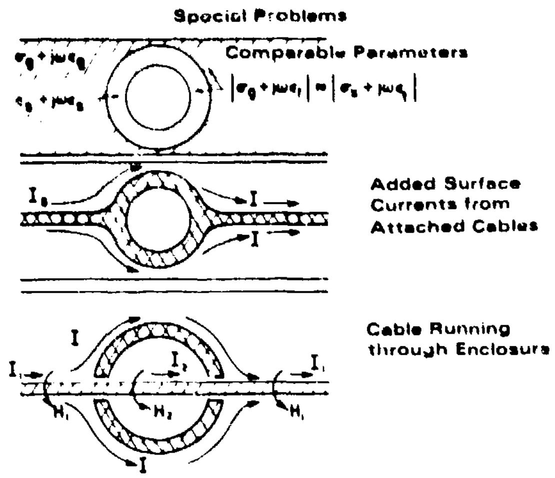
r = radial distance from the point in the enclosure at which the field strengths are to be determined to the center of the hole.



PENETRATION OF ELECTRIC AND MAGNETIC FIELDS THROUGH A CIRCULAR HOLE

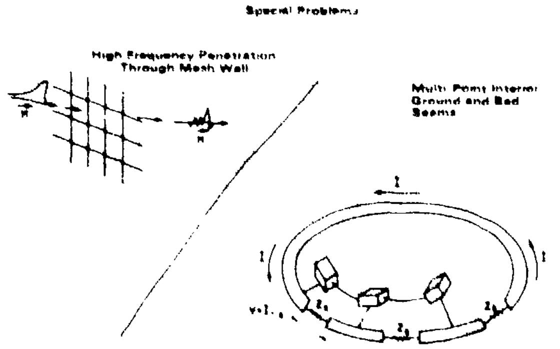
Special Problems

There are many special problems which we have not covered in this very brief discussion on shielding. As mentioned previously, there is a problem using the so-called rigorous or lumped-element approaches, wherein the parameters of the wall material are quite comparable to the ground parameters. Another case not considered, but one that must be considered in any shielding, is the collection of the additional current arising from attached cables as illustrated here. Obviously, we can get a lot more penetration if an exposed cable runs through an otherwise shielded enclosure.



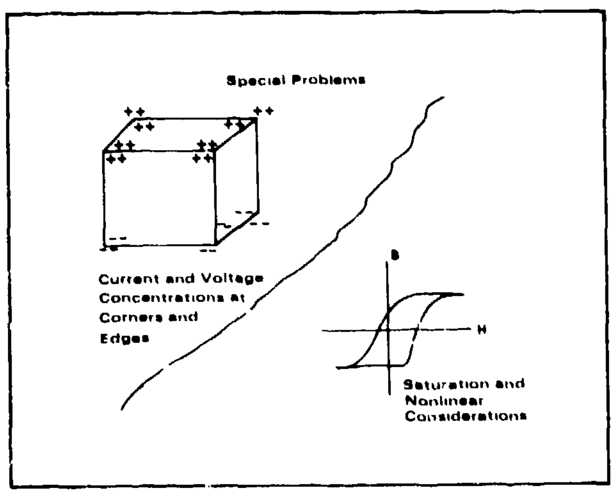
The mesh or loop-type shielding is valid only up to the point wherein the circumference of the loop is significantly smaller than the wavelength. Where the circumference of the loop is larger than the wavelength, various distorted types of penetration can occur, such as illustrated here.

We also have a multi-point ground problem. In many instances it is necessary to ground equipment to the walls of the enclosure at different points. If the inside wall of the enclosure is used as a return current path, serious problems may occur and may arise from lack of symmetry in the conductivity in construction of the wall enclosure. Specifically shown is the effect of seam resistance. If this enclosure is exposed to exterior fields, currents will flow on the outside of the enclosure. Due to the seam imperfections, a voltage drop occurs both on the outside and inside near the seams. This voltage drop can, in turn, then be injected into the circuits via a common ground or common impedance.



As mentioned in the previous section, field enhancement also plays a major role in coupling. The enclosure itself can also enhance the fields, especially near corners and edges.

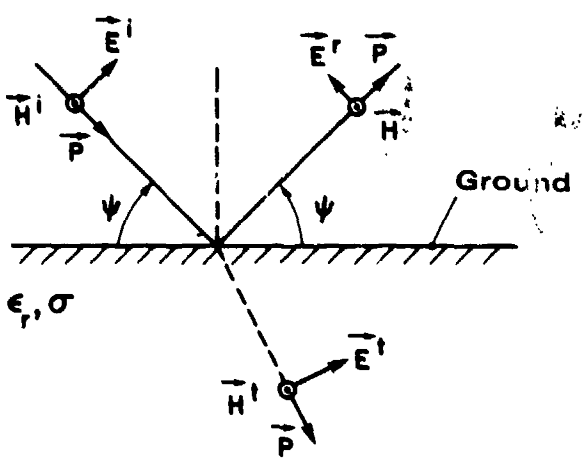
For certain situations, saturation and nonlinear effects of the wall material may be important. However, if the wall is designed of ordinary cold-rolled steel, copper, or aluminum and is of sufficient thickness to begin with, this, in general, is not a problem for the radiated or more distant fields.



Reflection and Transmission of EM Waves

To determine the response of an underground coupling structure to EMP, it is necessary to first calculate the sub-surface EMP fields in the absence of the structure. An approximate method for calculating the transmitted EM fields into the ground is to assume a plane surface and use ray theory in conjunction with appropriate boundary conditions. Consider now a vertically (E field in the plane of incidence) polarized EMP plane wave incident on the earth's surface at an elevation angle ψ.

Reflected and Refracted Waves at the Air-Ground Interface



It can be shown that the ground reflection coefficient for the vertically polarized case is given as:

$$R_V = \frac{E^r}{E^i} = \frac{(\varepsilon_r - jx)\sin\psi - \sqrt{(\varepsilon_r - jx) - \cos^2\psi}}{(\varepsilon_r - jx)\sin\psi + \sqrt{(\varepsilon_r - jx) - \cos^2\psi}}$$

where

$x = \frac{\sigma}{\omega\varepsilon_o}$

σ = ground conductivity

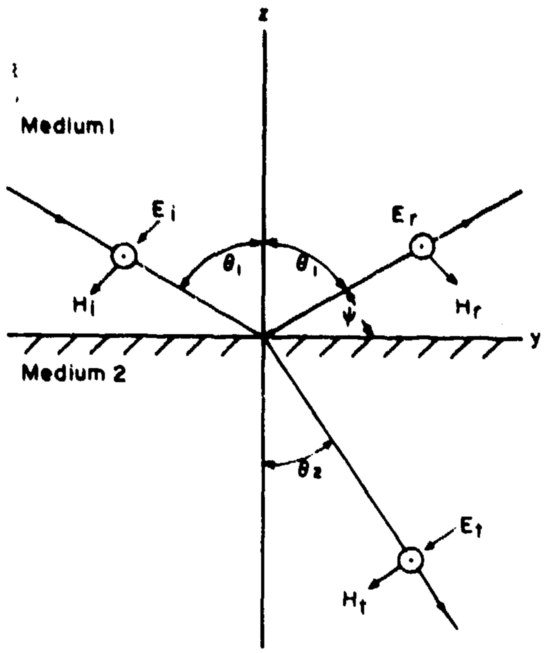
ε_r = relative dielectric constant of ground

ψ = elevation angle.

The transmission coefficient is defined as:

$$T_V = \frac{E^t}{E^i} = (1-R_V)\ \frac{\sin\psi}{\sqrt{1 - \frac{1}{\varepsilon_r - jx}\cos^2\psi}}$$

The geometry for the case of horizontal (E field normal to the plane of incidence) polarization is shown in the following figure.



REFLECTED AND REFRACTED WAVES--
HORIZONTAL POLARIZATION

The reflection coefficient for this case is given by:

$$R_H = \frac{E_r}{E_i} = \frac{\sin\psi - \sqrt{(\varepsilon_r - jx) - \cos^2\psi}}{\sin\psi + \sqrt{(\varepsilon_r - jx) - \cos^2\psi}}$$

and the transmission coefficient is

$$T_H = \frac{E_t}{E_i} = 1 + R_H$$

REFERENCES

"Electromagnetic Pulse Handbook for Missiles and Aircraft in Flight," EMP Interaction Notes 1-1, AFWL TR-73-68, Air Force Weapons Laboratory, Kirtland AFB, New Mexico.

"EMP Design Guidelines for Naval Ship Systems," Naval Surface Weapons Center/White Oak Laboratory, Silver Springs, Maryland (to be published).

"EMP Engineering and Design Principles," Bell Laboratories, Loop Transmission Division, Technical Publication Department, Whippany, New Jersey.

Jordan, E.C., Electromagnetic Waves and Radiating Systems, Prentice Hall, 1968.

Vance, E.F., "Prediction of Transients in Buried, Shielded Cables," Stanford Research Institute, SRI Project 2192.

King, R.W.P., Theory of Linear Antennas, Boston, Howard University Press, 1956.

King, R.W.P., Harrison, C.W., Jr., and Denton, D.H., Jr., "The Electrically Short Antenna as a Probe for Measuring Free Electron Densities and Collision Frequency in an Ionized Region," Journal of Research of the National Bureau of Standards, - Radio Propagation, Vol. 65, No. 4, pp. 371-383, July-August 1961.

Gooch, D.W., Harrison, C.W., Jr., King, F.W.P., and Wu, T.T., "Impedance of Long Antennas in Air and in Dissipative Media," Journal of Research of the National Bureau of Standards, - D. Radio Propagation, Vol. 670, No. 3, May-June 1963.

Toulios, P.P., et al, "Effects of EMP Environment on Military Systems," Technical Report No. E6114 under USAMERDC Contract No. DAAK02-68-C-0377, IIT Research Institute, Chicago, Illinois.

Toulios, P.P., Kaurs, A.R., "Antenna Users Manual for Linear Antennas in an EMP Environment," Volumes I and II, Technical Report E6239, under Contract No. DAAG39-72-C-0192, IIT Research Institute, Chicago, Illinois.

Baum, Carl E., "On the Singularity Expansion Method for the Solution of Electromagnetic Interaction Problems," Interaction Notes, Note 88, December 1971.

Vance, E., "DNA Handbook Revision, Chapter 11, Coupling to Cables," SRI, Menlo Park, Ca.

SECTION V

COMPONENT AND SYSTEM DEGRADATION

5.1 INTRODUCTION

Electromagnetic energy coupled into a system through deliberate antennas, via penetrations, or directly to internal circuit wiring due to apertures can degrade the system's performance. The degree of degradation is a function of many factors related to the normal operating mode of the system, the mission of the system, and the components utilized in the system. The purpose of this section is to present the mechanisms which cause performance degradation and provide the thresholds at which component degradation occurs.

Definitions

In order to understand the discussion that follows, it is necessary to clearly define the terminology that will be used in this and subsequent sections of the course. The terminology to be defined here relate to the ability of a system to perform its mission in the presence of an EMP.

To assess the impact of an electromagnetic pulse, or any other stimuli on a system's performance, the response of the system to the stimulus must be known. This response (in terms of degradation or upset) of a component, equipment package, discrete subsystem or system is termed the <u>susceptibility</u>.

SUSCEPTIBILITY

Susceptibility is defined as the response of individual components, equipment packages, discrete subsystems, or complete systems to a broad range of electromagnetic waveforms.

The conclusion that a system is susceptible does not mean that the system's performance is deteriorated, it means only that the system responds.

At the component level, susceptibility is usually determined empirically and may be expressed in terms of a threshold. For equipments, subsystems, and systems, certain coupling modes and component thresholds are implicit in the determination of the susceptibility characteristics. A knowledge of susceptibility permits a determination of the performance degradation for various conditions of exposure.

The reduced capability of a component, equipment, subsystem or system is termed the <u>degradation</u> of performance. The degradation may be determined by jointly considering the susceptibility and environment (stimulus) or more directly by experimental methods.

DEGRADATION

Performance degradation is the deterioration of some feature of a system in response to an undesired electromagnetic environment.

In some cases, some performance degradation can be tolerable. When the performance degradation exceeds the limits of satisfactory performance due to a stress, the system/component is considered <u>vulnerable</u> to that stress.

VULNERABILITY

System vulnerability relates the perform[illegible] of the mission within acceptable limits of degradation to certain hostile situations.

There are two types of degradation. They are:

1. Functional Damage

2. Operational Upset

Functional damage refers to permanent damage due to an electrical transient, while operational upset refers to temporary impairment due to an electrical transient.

FUNCTIONAL DAMAGE AND OPERATIONAL UPSET

If a system becomes permanently damaged due to a large electrical transient, it is said to have suffered *functional damage*.

The temporary impairment of a system's operation due to a smaller electrical transient is known as *operational upset*.

There are likely to be a variety of failure criteria. These may be broadly classed into three categories:

1. A catastrophic failure refers to a device which could not be expected to operate satisfactorily in any circuit.

2. A parametric failure refers to a device where parameter degradation has proceeded to a point where the circuit, although it continues to operate, will do so at reduced efficiency or lowered performance.

3. A state failure refers to an undesired change of state of a circuit.

FAILURE

There are likely to be a variety of failure criteria. These may be broadly classed into three categories:

1. ***catastrophic failure***
2. ***parametric failure***
3. ***state failure***

To define failure, therefore, it is necessary to consider the allowable degradation of performance of a component, circuit, equipment, subsystem or system. When this allowable degradation has been exceeded, failure has occurred.

History

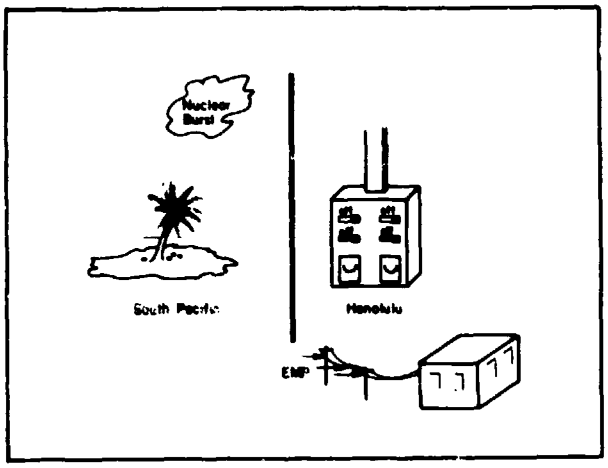
The first reported instance of EMP effects appearing in the open literature is from Electronic News, October 30, 1967,

"U.S. Seeks Answers to A-Blast Oddity."

The article reports:

> During the high-altitude nuclear tests in the Pacific in the early 1960's, "Hundreds of burglar alarms" in Honolulu began ringing, "circuit breakers on the power lines started blowing like popcorn."

Since there were no electrical storms in the area, it was concluded the EMP from a high-altitude nuclear test 500 miles away was the cause of these unusual occurrences.



Other failures noted during the atmospheric test era were associated with some of the very sensitive diagnostic instrumentation used to instrument the tests. Upset of timing circuits and some communication links were also evident. Little damage or upset of deployed systems was noted. This resulted because the systems deployed were primarily analog types and employed vacuum tubes for the most part which have been shown to be relatively hard components.

System tests, using non-nuclear simulation of the EMP, in recent years (the late 1960's to the present) have demonstrated that digital systems employing computers or computer type memories are potentially vulnerable to upset, and systems employing solid state components are potentially vulnerable to damage.

Devices which may be susceptible to functional damage due to electrical transients are:

1. Active electronic devices (especially high frequency transistors, integrated circuits, and microwave diodes).

2. Passive electrical and electronic components (especially those of very low power or voltage ratings or precision components).

3. Semiconductor diodes and silicon control rectifiers (especially those used in power supplies connected to public service or long cable runs).

4. Squibs, detonators, and pyrotechnical devices.

5. Meters, indicators, or relays.

6. Insulated RF and power cables (especially those running near maximum ratings and which are exposed to humidity or abrasion).

Potentially dangerous situations may also occur in the presence of explosive fuel vapors if an arc should happen to form. An example is that of rocket fuels containing premixed oxidizers.

Devices or systems which may be susceptible to operational upset due to electrical transients are:

1. Low-power or high-speed digital processing systems.

2. Memory units such as core memories, drum storage, buffers, via wiring.

3. Control systems for in-flight guidance.

4. Protection or control systems for the distribution of 60 or 400 Hz power.

5. Subsystems employing long integration or recycling times for synchronization acquisition or signal processing.

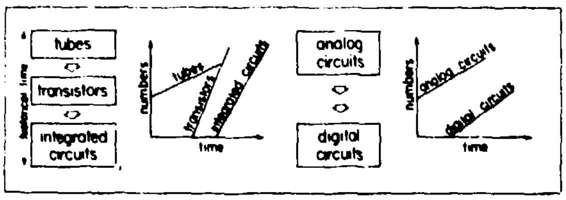
5.2 GENERAL DAMAGE AND UPSET CONSIDERATIONS

In order to assess EMP effects on a system, it is necessary to determine if the system will properly respond once the transient conditions have been damped out. These predictions require a knowledge of the threshold levels at which components fail (damage) and the threshold levels at which circuits temporarily malfunction (upset). The assessment problem is further complicated since for a component to fail or a circuit to upset, sufficient energy must reach the sensitive component or circuit. This energy collection and transfer problem is highly dependent on the system physical and electrical characteristics. It is these characteristics which determine the total available energy at the sensitive component, and the waveshape and fundamental frequency of the induced voltage and current at the sensitive component.

Trends

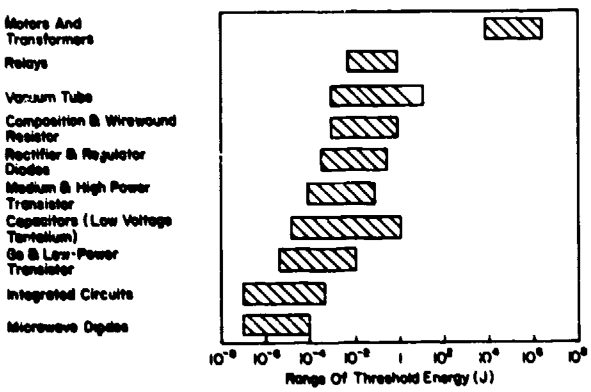
The modern trend in electronics is to more transient sensitive components and circuits. As vacuum tubes were replaced by transistors, and transistors by integrated circuits, electronic equipment has become more susceptible to damage from EMP. Even the passive elements now being incorporated into electronics have more susceptibility to damage than before. An example of this is the diffused and thin film resistors in integrated circuits.

TRENDS IN ELECTRONIC EQUIPMENT WHICH INFLUENCE EMP DAMAGE AND UPSET CONSIDERATIONS

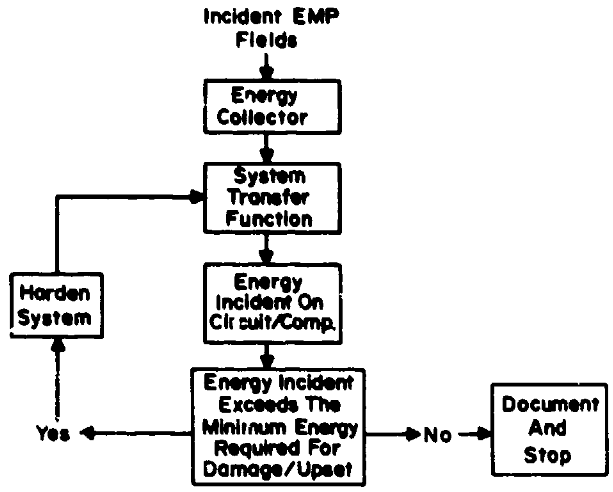


Widespread use of digital electronics equipment has placed emphasis on the importance of the effects of transient disturbances. This stems from the discrete, multi-static nature of digital equipment as compared to analog equipment which, although it may have many degrees of freedom, generally responds to transients by temporary excursions from steady-state operation conditions to which it returns after a time interval determined by circuit characteristics. With digital equipment, a momentary transient may cause the device to jump to a new state which is totally unrelated to the initial state, and from which it may of its own accord, never return after the transient has subsided.

A comparison of the susceptibility of common components or devices based on the threshold energy required for damage is shown in the figure. The threshold for upset is generally one to two orders of magnitude less than the minimum indicated for damage of the most sensitive components (i.e., 10^{-8} to 10^{-10} joules).



COMPARATIVE SUSCEPTIBILITY OF COMMON DEVICES

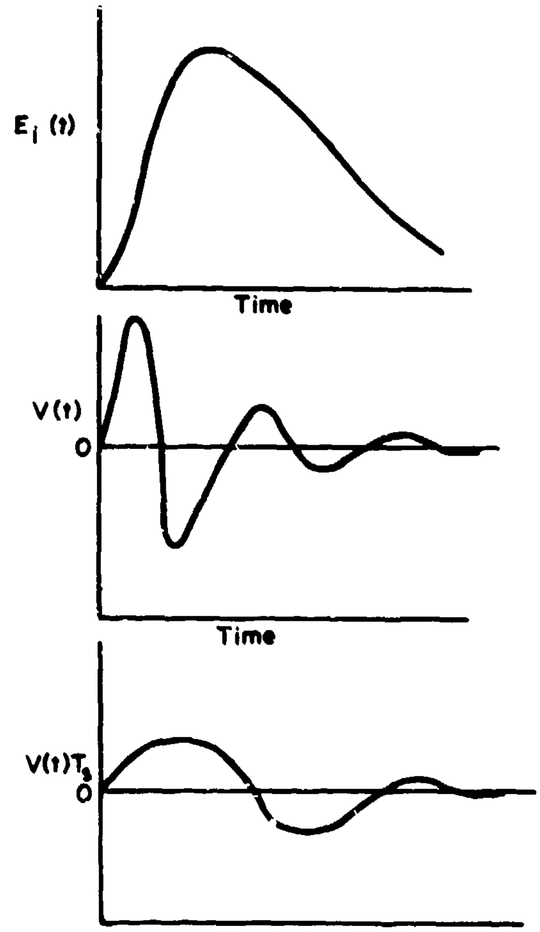


Consider the incident EMP fields from an exoatmospheric burst. For purposes of system assessment, the time waveform is characterized as a difference of exponentials providing a pulse with a rise time of a few nanoseconds and a fall time (first zero crossing) of nearly a microsecond. A transient with these rise and fall times will have very broad spectral content (approximately 10 kHz to 150 MHz).

System Configuration

The effects of EMP on a system depend not only on the energy collected by the system, but also on the nature of the circuits and components in the system. The EMP induced response in a circuit or component depends on the relationship between the time/frequency domain of the incident EMP fields and the time/frequency response of the system of interest.

The energy collectors (cables, antennas, apertures in shields, etc.) associated with real world systems do not respond uniformly over the entire frequency spectrum. Rather they respond most strongly at their own fundamental resonances which generally produce a coupled voltage waveform with the characteristics of a damped sine wave. This coupled waveform may or may not be further modified prior to arrival at the sensitive portion of the circuit from either a damage or upset viewpoint. This depends entirely on what frequency selective or amplitude limiting circuitry may be employed.



It is obvious from this discussion that the effects of EMP on a system are highly dependent on the system characteristics. To develop a simple failure model and derive comparative damage constants for components empirically, a uniform test procedure was necessary. A conservative model has been developed utilizing square wave pulses and testing components out of circuits to empirically determine the damage constants of components. The test pulse duration is related to the frequency of the damped sinusoid by the following equation:

$$t = \frac{1}{5f}$$

t = test pulse duration

f = frequency of damped sinusoid

The damage constant (damage level) data presented later was obtained in this manner.

Upset is primarily a system or circuit related problem rather than a component problem. It is concerned with changing state of digital logic, memory erasure, etc. Whether or not an interfering waveform will result in the introduction of errors (change of state of flip-flops or gates) depends on the characteristics of the waveform (rise time, duration, and amplitude) and the circuit (bias conditions, type of logic, etc.). Memory erasure depends on the driving current into the memory cores, or recording heads. Capacitive discharge (exponential pulses) were used to determine the minimum energies to cause the resultant action presented later as a rule of thumb guide for a comparison with the damage levels presented. These data cannot be used as other than a rule of thumb guide, however, due to the degree of waveform/circuit dependence.

5.3 COMPONENT FAILURE

A number of damage mechanisms have been observed for electronic components subjected to electrical transients. Some of these are

- Dielectric breakdown
- Thermal effects
- Interconnection failures

The voltage at which dielectric breakdown occurs is a function of the material and the thickness of the material. Breakdown can occur in all types of insulating layers if the voltage stress is high enough and applied for a sufficient time (pulse duration). In the case of insulators, this generally occurs as surface breakdown. In the case of electronic components, it may occur as surface breakdown or internal breakdown.

Thermal effects result from the dissipation of energy in the component due to excessive current flow. This is a major cause of semiconductor junction failure and resistor burnout. Thermal effects may also be responsible for such failures as spot welding of relay contacts, and detonation of electro-explosive devices employing bridge wires.

Interconnection type failures result from the induced electrical transients increasing the temperature sufficiently to cause melting of metal surface connections, beam leads on integrated circuits, and the wire in wire-wound resistors.

These effects can result directly from the EMP induced voltages and currents or indirectly due to power follow through. Power follow through occurs where the EMP induced transient serves to trigger a particular effect and where sufficient energy can then be supplied from other sources (such as transmitter outputs, power supply, etc.) to cause permanent damage.

These effects are discussed as they apply to various components in the following paragraphs.

Semiconductor Device Failure

The initial understanding of semiconductor device failure is best obtained by considering a single P-N junction. Subsequent extrapolation of the phenomena to multijunction devices is relatively straightforward.

The principal failure mechanisms for a single P-N junction are:

PRINCIPAL FAILURE MECHANISMS FOR A SINGLE P-N JUNCTION

1. For Reverse Voltages
   (a) Surface Breakdown Around the Junction
   (b) Dielectric Breakdown
   (c) Internal Breakdown Through the Junction Within the Body of the Device
2. For Forward Voltages
   (a) Internal Breakdown in the Body of the Device

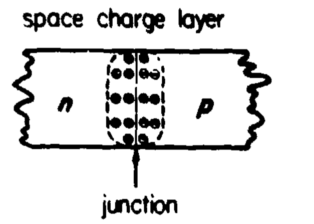
These failure mechanisms are discussed in the following paragraphs.

Surface Effects

The destruction mechanism of a surface breakdown is usually to establish a leakage path around the junction, thus nullifying the junction action. The junction itself is not necessarily destroyed and re-etching the surface can return the junction to normal operation. The problem of theoretically predicting surface breakdown is difficult since it depends upon many parameters such as geometrical design, doping levels near the surface, lattice discontinuities on the surface, and general surface conditions.

It is well known that the surface of a P-N junction influences the electrical characteristics of the semiconductor device. For the junction perpendicular to the surface, surface breakdown can be explained as a localized avalanche multiplication process caused by narrowing of the junction space charge layer at the surface.

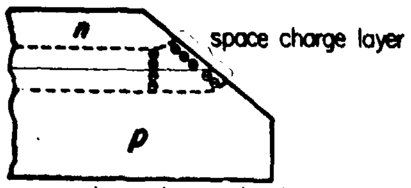
SURFACE BREAKDOWN AROUND THE JUNCTION



space charge layer in reverse voltage mode

It has been shown experimentally that the electric field is altered by the contour of the semiconductor surface in the vicinity of the junction and proper contouring of the surface of a P-N junction results in a lower potential gradient at the surface.

SURFACE BREAKDOWN AROUND THE JUNCTION



space charge layer with bevelled p-n junction

The space charge layer is then spread over a greater surface distance than it would occupy if the surface edge were perpendicular to the junction. For the contoured surface, the maximum electric field at the surface will be less than that within the body of the device. By properly contouring the surface, the peak electric field there can be reduced to a fraction of that in the interior of the device. Hence, it is possible to build junctions which exhibit body breakdown prior to surface breakdown, thus eliminating this type of breakdown as a principal failure mechanism.

Since avalanche breakdown occurs more easily on a surface than within the body of a device, and since it is strongly field dependent, a reduction in electric field on the surface of a P-N junction device is also desirable from this standpoint.

Dielectric Breakdown

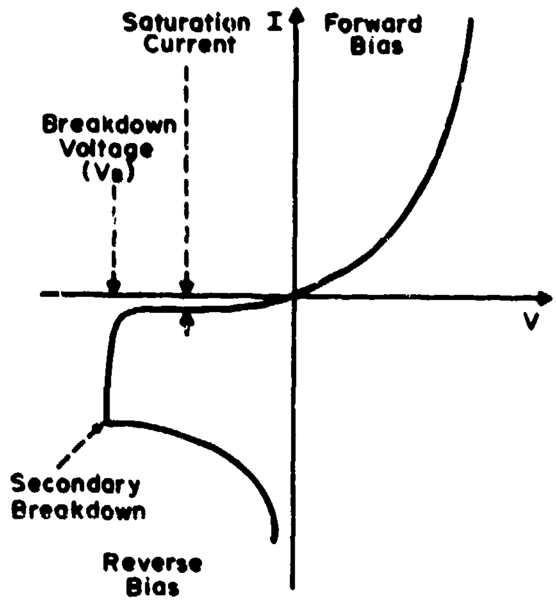
Junction failure due to dielectric breakdown is a result of a large avalanche current which forms a path for an arc discharge to occur. This can result in a puncture through the junction with an actual pinhole being formed. Usually a junction short is caused.

Most semiconductor dielectric layers are thick enough to withstand severe electrical transients. Thin layers, such as those found in fast switching devices (insulated gate field effect transistors), can breakdown at dc voltages ranging from 30 to 200 volts. Dielectrics can withstand higher transients and ac voltages, but if the transient persists long enough for avalanching to occur (about 1 microsecond), they can still be damaged.

Internal Junction Breakdown

In internal body breakdown, the destruction mechanism apparently results from changes in the junction parameters due to localized high temperatures within the junction area. These temperatures can be of such magnitude that alloying, or diffusion of the impurity atoms occurs to such an extent that the junction is either totally destroyed or its properties drastically changed. The current may be sufficiently high and localized to cause melting at hot spots within the junction. Such action can result in a resistive path(s) across the junction which develop after resolidification of the melt at the junction. The primary effect on device operating characteristics is manifested as a decrease in diode breakdown voltage and an increased leakage current, while in transistors, decreased gain and increased junction leakage currents are observed.

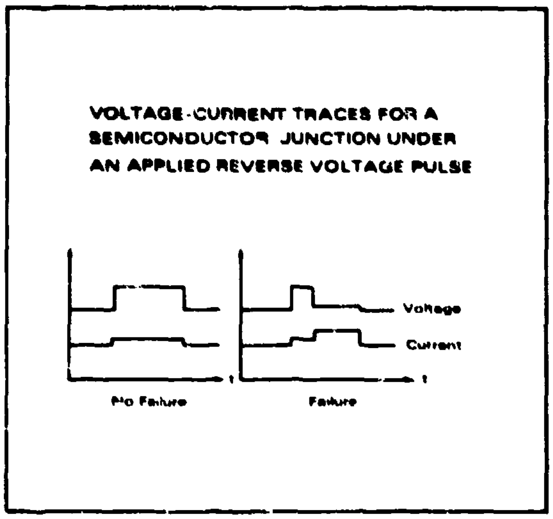
The major cause of semiconductor failure occurs under reverse bias conditions. This failure mechanism is termed "secondary breakdown." The voltage-current curve for a P-N junction indicates that for low reverse voltages the device conducts only a very small current. As the reverse voltage increases, breakdown (V_B) occurs with a resulting increase in current flow. A major portion of the energy during breakdown is dissipated in the junction, since the junction is reversed bias, resulting in heating of the junction. This results in a one type of second breakdown termed "thermal second breakdown."



VOLTAGE-CURRENT RELATIONSHIP OF P-N JUNCTION

Thermal second breakdown physically is a local thermal runaway effect at the junction induced by severe current concentrations within the device which are a function of the biasing conditions, excessive junction fields, and material defects. One reference considers second breakdown as a filamentation phenomenon which occurs in three stages: nucleation of the filament, growth of a relatively broad filament across the high resistivity region, and growth of a second filament interior to the first wherein material is in a molten stage. The first two are non-destructive. The third involves the formation of a melt channel which results in irreversible device degradation. Under reverse bias, nucleation of a current filament starts at a localized region of high current density in the junction. More than one filament can form, depending upon the conditions of excitation and the device geometry.

The voltage across and the current through the junction as a function of time when a high amplitude transient is applied as a reverse bias usually exhibits a high voltage, low current characteristic. When thermal second breakdown occurs, the case when the junction fails, the voltage suddenly decreases and the current rapidly increases.



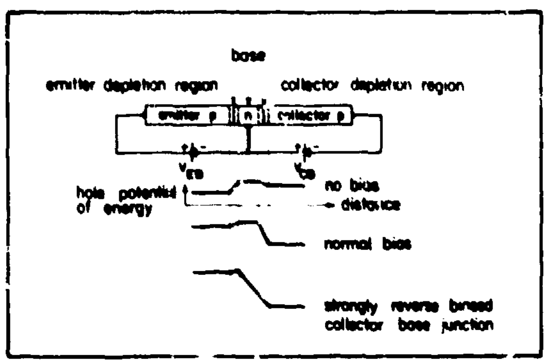
It has been observed that the occurrence of thermal second breakdown, which is an energy dependent process, represents the point of incipient permanent damage for semiconductor devices at submicrosecond pulse conditions. That is, once a thermal second breakdown is initiated with some additional amount of energy being further dissipated in the device, a permanent damage condition results. For most semiconductor devices investigated, device degradation after second breakdown was a result of junction damage, i.e., realloying of the material or the formation of a melt channel. However, for some devices, a low voltage-high current mode of operation was observed due to the migration of contact metalization through the junction thus forming a metallic short. Depending upon the device, either of these phenomena could possibly occur initially, thus precipitating device failure.

Another reverse bias failure mode which has been observed on occasion in transistors is that of current mode second breakdown. Basically, the effect is initiated by relatively high material current densities under the emitter during collector to base junction reverse pulsing resulting in a forward bias on a portion of the emitter. When this bias becomes sufficiently high, the device becomes unstable and is switched to a low impedance, low sustaining voltage mode of operation.

The occurrence of the current mode second breakdown phenomenon in itself generally has not been observed to result in permanent damage. However, if the resulting low impedance currents are not sufficiently limited, then local hot spots form and the resulting failures are produced in a manner similar to that of thermal second breakdown.

The forward biased junction vulnerability to pulsed electrical energy can be understood by considering some of the basic concepts associated with thermal second breakdown. That is, device degradation is a direct result of essentially similar melting and realloying reactions at various current constriction sites within the junction. The significant fact here is that due to the relatively lower junction voltage at forward bias conditions, a correspondingly higher current than that for the reverse direction is required to reach the critical failure energy. This larger current, in turn, results in a significant voltage drop being produced in the bulk material. Hence, a higher energy input is generally required as far as the device terminals are concerned, thus producing much of the apparent decrease in failure sensitivity observed experimentally. Again, since a relatively low initial impedance condition exists, the dramatic switching associated with thermal second breakdown would not generally be observed.

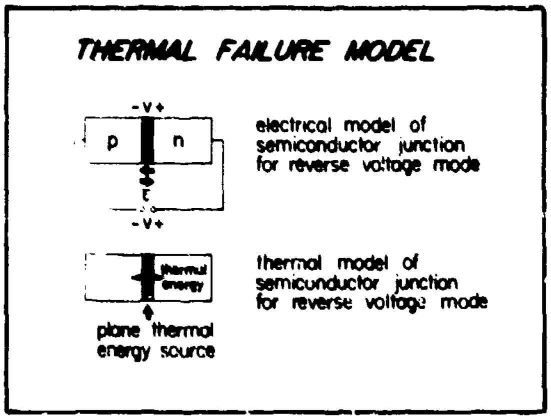
In addition to the single junction mechanisms, another failure mechanism in transistors is possible due to its multijunction nature. This mechanism is called punch-through. The width of the depletion region at a reverse-bias junction will increase as the voltage across the junction increases. Since the collector-base junction of a transistor is usually reverse biased and of small width, it is possible for the depletion region to extend throughout the width of the base which effectively results in a short circuit. Under these conditions, the resulting current may be sufficiently large to damage the junction.



Thermal Failure Model

Many different microscopic mechanisms may contribute to semiconductor failure. However, most of these mechanisms have been found to be linked primarily to the junction temperature. Therefore, in most cases, the treatment of the problem can be reduced to a thermal analysis.

The worst case as far as achieving high temperatures in the junction is when one considers that all of the power dissipated in the device occurs in the junction. This corresponds to the situation where a high-voltage pulse of reverse polarity is applied to a junction with a high reverse voltage breakdown. When the avalanche breakdown occurs, almost all of the applied voltage is dropped across the junction and only a small percentage is dropped across the bulk material (except for a very short pulse on the order of 10 to 100 nanoseconds or less where the high current required for failure causes more voltage drop across the bulk).



The thermal model for pulses between 100 nsecs and 1 millisec is based on the following assumptions:

- The heat is generated at the junction
- The junction is planar
- The silicon material on either side of the depletion layer of the junction extends out infinitely.

Then the one dimensional heat equation can be used to solve for the junction temperature.

$$\frac{\partial}{\partial x} K \left(\frac{\partial T}{\partial x}\right) - PC_p \frac{\partial T}{\partial t} = 0$$

where

K = the thermal conductivity of silicon (W/Cm°K)

T = the temperature (°K)

x = position measured from the planar heat source in Cm

P = density of the silicon (gm/Cm³)

C_p = specific heat of silicon (J/gm°K)

If a square pulse of electric power is applied to the junction and all of the electrical power is transformed to heating, the maximum temperature of the junction is given by:

$$T_m = T_i + \frac{P}{A} \frac{1}{\sqrt{\pi KPC_e}} t^{\frac{1}{2}}$$

where

T_m = the maximum junction temperature

T_i = the initial junction temperature

P = the electrical pulse power applied to the junction

A = the area of the junction

t = the pulse width

Rewriting the equation and taking the logarithm of both sides yields an equation which plots as a straight line with a -1/2 slope on log-log paper.

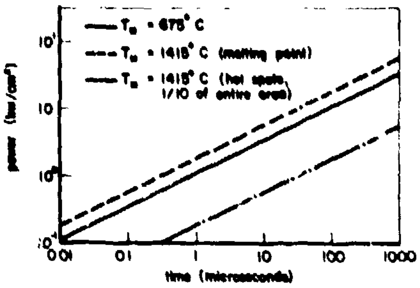
THERMAL FAILURE MODEL

$$P_m = A\sqrt{\pi k \rho C_p}\,[T_m(0) - T_i(0)]\,\frac{1}{\sqrt{t}}$$

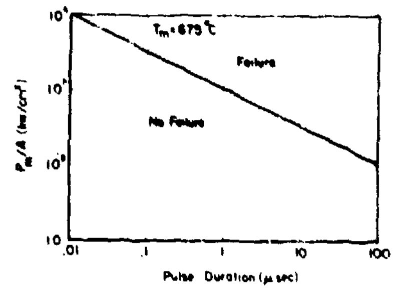
$$\log_{10} P_m = \log_{10} K - \tfrac{1}{2} \log_{10} t$$

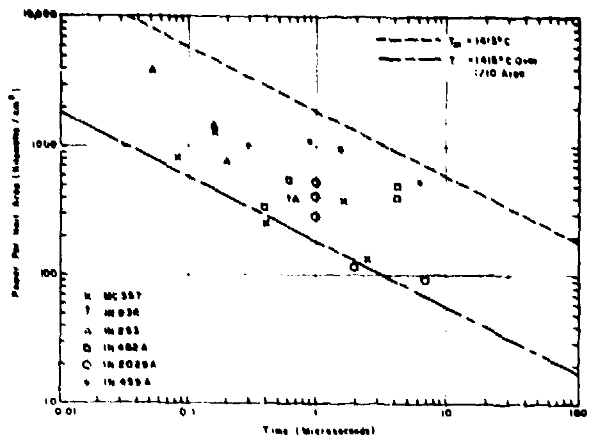
This model allows for determing the peak pulse power as a function of pulse duration if the junction parameters and failure temperature of the junction are known. The theoretical failure curve shown is for a silicon junction with the failure temperature of 675° assumed.

REVERSE VOLTAGE FAILURE CURVES FOR SILICON BASED ON TOTAL ENERGY DELIVERED TO THE JUNCTION

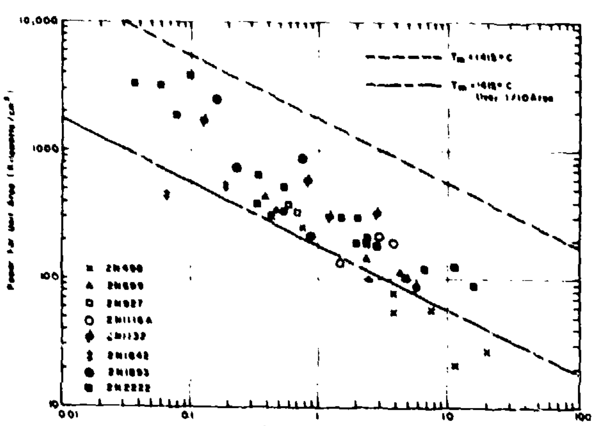


THEORETICAL FAILURE CURVES FOR SILICON JUNCTIONS FOR REVERSE VOLTAGE MODE





EXPERIMENTAL DATA FOR DIODES



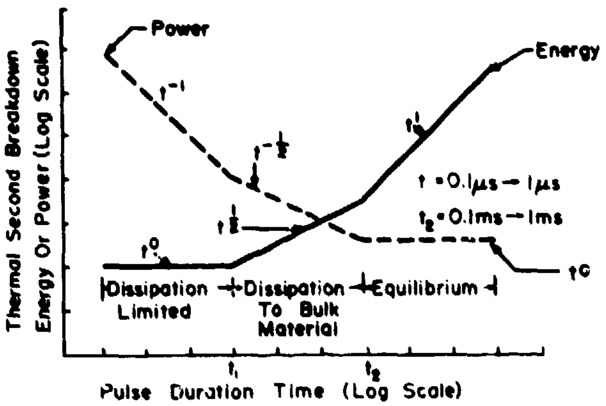
EXPERIMENTAL DATA FOR TRANSISTORS

The exact value of the final temperature at which component failure occurs is open to question. Usually, 675° is considered as the temperature at which silicon ceases to be intrinsic. That is, at this temperature, diffusion of impurity atoms across the junction can be expected which tends to nullify junction action. In general, the exact temperature is a function of the doping level and other factors. The melting temperature (1415°C) of silicon, of course, leaves no doubt as to junction destruction. Generally, it would be expected that significant junction degradation occurs before the melting point is reached.

Another problem encountered in attempting to correlate experimental data with the theoretical model, results from the fact that under reverse bias conditions, the current density across the plane of the junction is not uniform. Constriction of the current to a few (hot spot) sites effectively reduces the available junction area. Calculation of the reduced area is beyond the capabilities of a reasonably simple analytical model. Partial experimental results yield an average effective junction area reduction of 20 to 30 percent for some components. A curve for a 10% effective area is shown in the figure with a corresponding reduction in the required failure power density by a factor of ten. Most experimental data appears to be bracketed by these theoretical curves. Typical data is shown for both diodes and transistors in the accompanying figures.

The discussion up to this point is for applied pulses with durations between 100 ns and 1ms. In this case, the cause of junction failure is due to localized heating with the heat source at the junction. Using the one dimensional heat flow equation results in the $t^{-1/2}$ relationship as shown. For shorter or longer pulses, the one dimensional heat flow equation and the plane heat source at the junction do not apply.

For short pulse widths (<100 Ns), a constant energy condition independent of pulse width prevails for the initiation of junction failure. From physical considerations, this is the required energy input to a volume of material (the current constriction site) in order for that volume to achieve an increase in temperature under adiabatic conditions. A one-half power of time dependence is obtained for longer pulse widths, which is indicative of heat loss from the (constriction site) volume to its surrounding medium. The direct time dependence for energy found at the longer pulse widths (signifying a constant power input) is indicative of thermal equilibrium resulting in a steady-state temperature at the center of the volume. This constant power level approaches the manufacturer's CW rating for the device as the pulse width becomes very large (>1 ms).

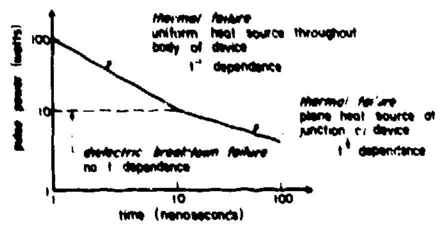


TYPICAL FAILURE LEVEL RELATIONSHIPS FOR THERMAL SECOND BREAKDOWN IN SEMICONDUCTOR JUNCTIONS

For failure due to surface breakdown, no τ dependence results and the failure level is dependent only on the pulse power, or equivalently, the failure level is power or voltage-dependent and not energy-dependent. For example, the failure of microwave diodes for 2-20 nanosecond pulses has been found to be dependent on peak power rather than energy.

Other devices have also been found to be voltage sensitive. This voltage sensitivity seems to occur sometimes across the surface, i.e., surface flashover; and for higher voltage pulses by punchthrough. This voltage sensitivity is a rise-time effect and can occur in some devices with a longer pulse (>30 nanoseconds) provided that the mechanism responsible for the voltage sensitivity occurs before any other effect, viz., thermal failure.

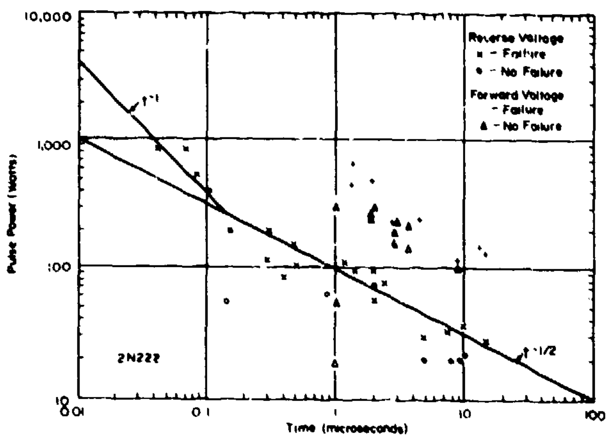
JUNCTION FAILURE FOR PULSE LESS THAN 30 NANOSECONDS



Verification of Wunsch Thermal Model

A number of semiconductor devices (primarily diodes and transistors) have been experimentally tested at various laboratories. Empirical relationships have been derived from the experimental data for the component burnout energy or power as a function of pulse width. It has been shown experimentally that the proportionality between energy or power for component burnout varies as a function of incident pulse width as shown previously.

Typical of data obtained, is the following curve for the 2N2222 transistor showing the single pulse failure power. The change in slope of the best straight line fit to the experimental data for the shorter width pulses is evident.



FAILURE OF THE BASE-EMITTER JUNCTION OF A 2N22 TRANSISTER FOR FORWARD AND REVERSE POLARITY VOLTAGE PULSES

Similar data have been obtained for other semiconductor devices by a number of experimentalists. In general, it has been found that the required power level for failure is approximately proportional to the device junction area for areas of up to 10^{-2} to 10^{-1} cm^2. (For larger areas, the failure level becomes more area independent). Because of this proportionality, it has become commonplace to plot power-to-failure per unit-of-junction area versus pulse width. This normalization of the power failure curve allows the convenient comparison of data for more than one device type on one graph.

The general relationship between threshold failure power as a function of pulse width may be considerably more complicated than indicated by the theoretical model. In general this relationship can be obtained experimentally and expressed in the form

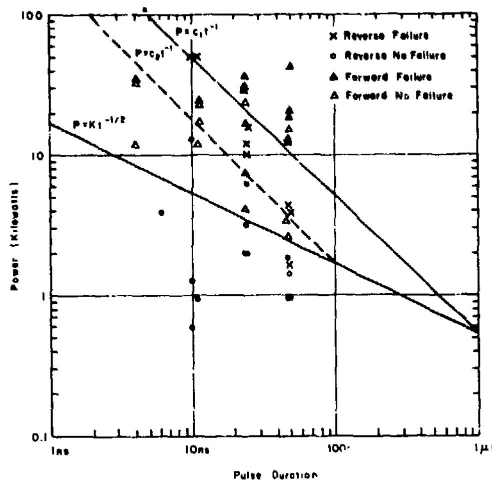
$$P_{TH} = At^{-B}$$

where A and B are determined by a least squares fit to the data. Much of the data fits the Wunsch model ($B = \frac{1}{2}$) for simple junction devices fairly well (i.e., within the experimental data spread of approximately 1 order of magnitude).

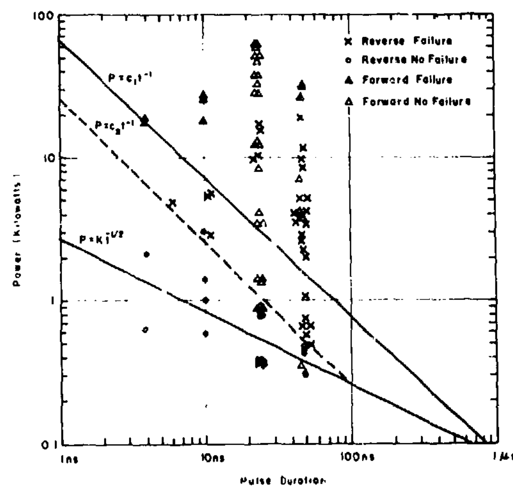
For pulse widths less than on the order of 100 nsec, the preceding relationship ($B = \frac{1}{2}$) between power failure level and the pulse width does not hold. In this regime, uniform heating of the semiconductor bulk material takes place leading to a modification of the power failure relationship to

$$P \approx C\, t^{-1} \text{ watts}$$

This form of relationship is plotted in the following figures for 2N336 and 2N2222 transistors, respectively. Two curves with slope proportional to t^{-1} are drawn for comparison with the trend of the experimental data points. These are the $P = C_1 t^{-1}$ curve whose intercept point has been adjusted so as to intersect the $P = Kt^{-\frac{1}{2}}$ curve at t = 1 μsec and the $P = C_2 t^{-1}$ curve which intersects at t = 100 nsec. Inspection of the curves shows that the best fit to the data occurs for different intercept point curves for each device type. In general, the optimum intercept point is a function of device thickness and will tend to move to larger time values for the larger devices.



EXPERIMENTAL DATA FOR BASE EMITTER JUNCTION FAILURE OF A 2N336 TRANSISTOR



EXPERIMENTAL DATA FOR BASE EMITTER JUNCTION FAILURE OF A 2N2222 TRANSISTOR

The Damage Constant - K

The power failure threshold (P_{TH}) is different for different classes of devices and devices within a class. The power failure threshold P_{TH} is defined as

$$P_{TH} = At^{-B}$$

where

A = damage constant based on the device material and geometry

B = time dependence constant.

The constants A and B may be determined empirically for every device of interest by the least squares curve fit to the data. The theoretical model, however, has a time dependence constant of $B = \frac{1}{2}$. In the mid-range of pulse widths, say from approximately 100 ns to 100 μs, the empirical data for a wide range of devices fits the model within the experimental data spread. In order to be able to directly compare device susceptibility for various pulse widths, and since the data fit fairly well, they are fit to the theoretical equation:

$$P_{TH} = Kt^{-\frac{1}{2}} \text{ Kw/Cm}^2$$

where

t = the pulse width in microseconds

K = the Wunsch model damage constant in kW-(microsecond)$^{\frac{1}{2}}$

It is convenient to express K in these units since the numerical value of K is then equal to the power necessary for failure when a one microsecond pulse is applied to the junction.

In adapting the model for diodes and transistors, Wunsch used experimental data for similar type devices with known junction areas to obtain the best curve fit. The results were:

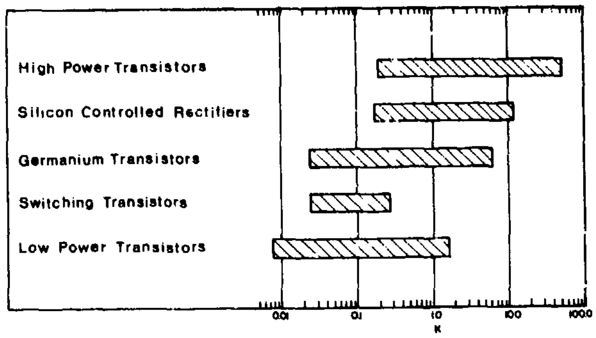
For diodes:

$$\frac{P_{TH}}{A} = 550\ t_o^{-\frac{1}{2}}; \text{ or } K = 550A$$

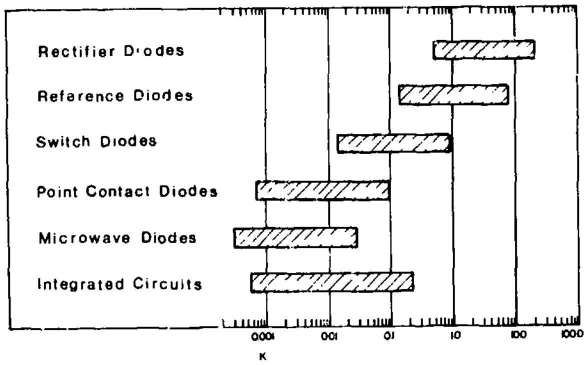
For transistors:

$$\frac{P_{TH}}{A} = 470\ t_o^{-\frac{1}{2}}; \text{ or } K = 470A$$

Knowing the junction area, therefore, provides the value for the damage constant, K. Typical range for K for various semiconductors are shown in the following figures. Multiplication of this factor by $t^{-\frac{1}{2}}$ will yield the pulse power threshold.



RANGE OF PULSE POWER DAMAGE CONSTANTS FOR REPRESENTATIVE TRANSISTORS AND SCR'S



RANGE OF PULSE POWER DAMAGE CONSTANTS FOR REPRESENTATIVE SEMICONDUCTORS

The actual value of K for a specific device may be found, of course, experimentally. Data on some specific diodes and transistors are presented in the tables.

DAMAGE CONSTANTS
DIODES

Type	K	Type	K
IN547	12.1	IN914	0.85
IN625	0.164	IN936	0.14
IN625A	0.045	IN936A,B	7.0
IN645	2.8	IN968B-979B	1
IN660	0.44	IN1200, 1201	62.32
IN662	0.29	IN1317A	0.19
IN689	1.1	IN2808	249
IN702	1	IN2970B	15.0
IN709	0.78	IN3017B	1.9
IN719A	0.1	IN3064	0.02
IN746A-IN752A	1.1	IN3821	1.947
IN754A-IN758A	0.63	IN3976	132
IN761,2,3	1.8	IN4370A	0.625
IN821	0.577	IN4823	0.208
IN823	1.8	D4330	0.001

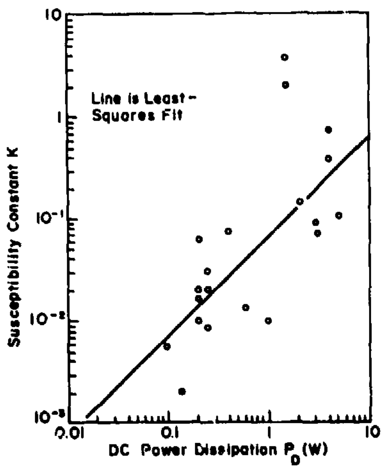
DAMAGE CONSTANTS TRANSISTORS

Type	K	Type	K
2N43,A	0.28	2N1480	5.5
2N43,244	0.05	2N1481	2.2
2N274	0.0076	2N1564	0.56
2N339	2.0	2N1602	0.40
2N404	0.05	2N1701	4.5
2N424A	10.0	2N1890	0.27
2N525	0.3	2N1916W	2.22
2N656	0.2	2N2035	3.633
2N687	11.7	2N2218A	0.264
2N717	0.13	2N2219	0.3
2N910	0.218	2N2219A	0.264
2N1039	1.4	2N2222,A	0.1
2N1154	21	2N2223,A	0.21
2N1212	13.129	2N3819	0.22
2N1309	0.087	2N3907	0.165

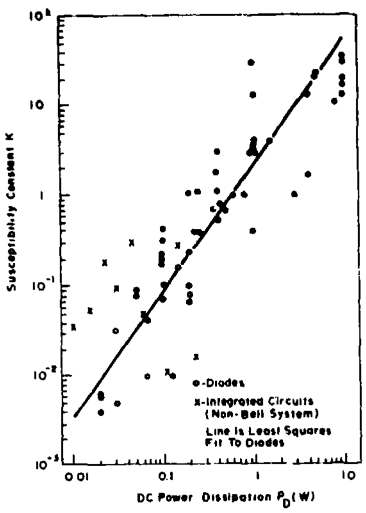
It should be noted that with a reasonable accuracy, it is possible to calculate the value of K from procedures based on a knowledge of either the semiconductor junction area, its thermal resistance, or junction capacitance. Discussions of these procedures are available in other references. The utility of these procedures lies in the fact that one or more of these junction parameters are normally available from manufacturer's data sheets.

It is interesting to note the relation between the susceptibility (damage) constant K and the dc power dissipation capability for various devices. As can be seen from these curves, there is a direct correlation between the dc power dissipation and the damage constant.

RELATION BETWEEN SUSCEPTIBILITY CONSTANT AND DC POWER DISSIPATION CAPABILITY: TRANSISTORS



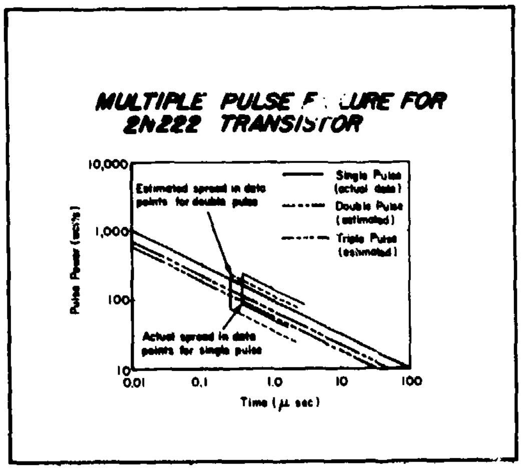
RELATION BETWEEN SUSCEPTIBILITY CONSTANT AND DC POWER DISSIPATION CAPABILITY: DIODES AND INTEGRATED CIRCUITS



Effect of Multiple Pulses

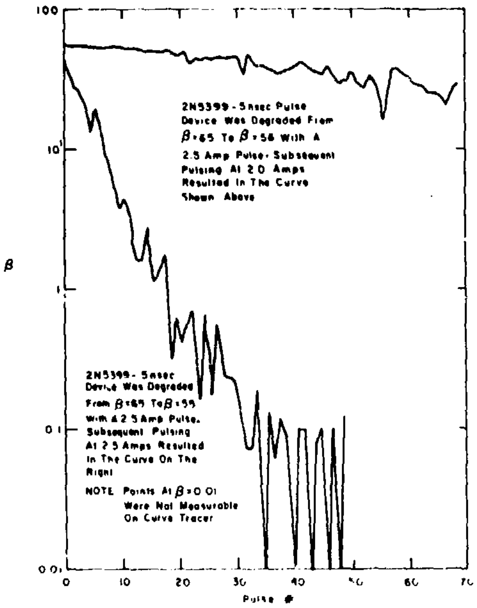
When a system is exposed to EMP, the components of that system will normally experience a transient that has a damped sinusoidal waveshape. This waveshape may be approximated in the laboratory by a series of single pulses to investigate the cumulative effects with or without a cooling period between the pulses.

The actual input power level at which a junction fails does not seem to decrease significantly with multiple exposure. This is illustrated for the case of the failure threshold for a 2N2222 transistor as a result of single, double, and triple pulse exposure. The interval between pulses was short enough that no junction cooling would take place between pulses. The energy in deriving these curves (double and triple pulse) which results in increasing the junction temperature was estimated as 2 and 3 times the single pulse energy.



It can be seen that the difference between single-pulse and triple-pulse failure thresholds is less than the experimental spread in data points for single-pulse failure. Hence, failure thresholds which are determined from single-pulse tests are usually adequate approximations for assessing the vulnerability of semiconductors to EMP.

The previous discussion was concerned with the heating effect of multiple pulses affecting the damage threshold level. Next, the cumulative effects of repeatedly pulsing a transistor at and below the damage threshold respectively will be considered. In this case, the pulses are far enough apart so that cumulative heating effects are negligible.



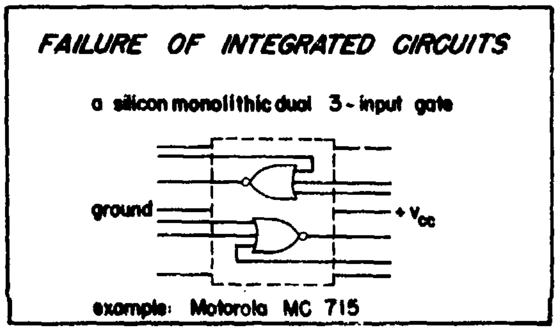
GAIN DEGRADATION FROM CURRENT PULSES FOR TWO CONDITIONS

The figure presents the results of multiple pulsing of a transistor for two different conditions. In each case, the initial pulse applied exceeded the degradation threshold of the transistor resulting in a reduction of the transistor β from 65 to 56. In the lower curve, this same pulse level was utilized for all subsequent pulses. The curve shows that the transistor gain increases and decreases erratically with each subsequent pulse. This indicates that multiple conducting filaments are probably formed across the junction when the device is damaged and subsequent pulses either add to the number of these filaments thus further decreasing the gain, or they cause a re-opening of existing filaments thereby producing an increase in gain.

The upper curve was obtained by pulsing the device at 0.8 times the damage threshold after the initial pulse degradation as in the previous case. Continued degradation occurs for each subsequent pulse. Comparison of the two curves indicates that reducing the amplitude of the applied pulses after damage initiation merely reduces the rate at which additional damage occurs.

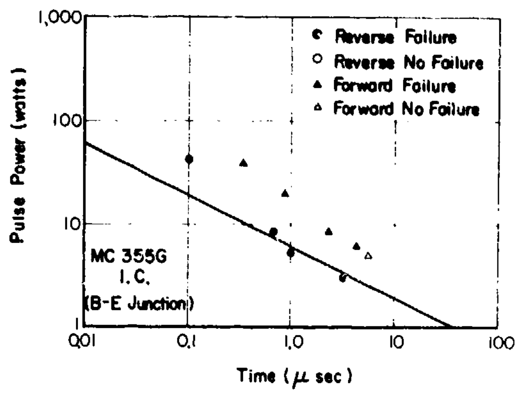
Integrated Circuit Failure

Integrated circuits (IC's) employ a large number of junctions on a single clip. Depending on the type of IC, it is possible that the state of the logic (digital logic circuit) voltage at one input will affect the burnout level when a pulse is applied to another input. This was investigated in the case of the MC715 circuit. It was determined for this particular device that the gates fail independently and the state of the logic voltage at the other inputs did not affect the burnout level. This device has isolated transistors forming the 3-input gate. This would not necessarily be true if the input circuits were not isolated.



The input and output leads of an IC are more susceptible to transient damage than the positive battery lead. This is not an unexpected result since a transient entering via the positive battery lead would be distributed over a number of P-N junctions in the device.

The pulse power required for damage of an MC 355 G integrated circuit (B-E junction) is shown in the following figure. The line is the thermal failure model curve with a slope of -1/2 fit to the data. As in the case of discrete devices, for pulse widths between 0.1 and 10 microseconds, the thermal failure model is appropriate.



The following table shows the experimental power failure threshold for 15 devices when exposed to a 1 μsec pulse. Data are given for the input lead, output lead, and battery lead. The minimum power for damage on any lead is the controlling threshold value and may be defined as the failure threshold power. On this basis, a K value can be assigned and for the devices indicated ranges from 1.1×10^{-2} to 5.0×10^{-1}. These K values are within an order of magnitude of those for diodes with the same dc power dissipation capability.

INTEGRATED CIRCUITS EXPERIMENTAL DAMAGE POWERS*

DEVICE	TYPE	EXPERIMENTAL FAILURE POWER (W)		
		INPUT LEAD	OUTPUT LEAD	BATTERY LEAD
Fairchild 9930	Dual 4-input gate	230	92	210
Signetics SE 8401	Quad 2-input Nand gate	72	47	390
T. I. 946	Quad 2-input Nand gate	15	20	275
Sylvania SG 140	Quad 2-input Nand gate	84	67	210
Motorola MC 3010	8-input gate	470	300	1800
Radiation, Inc. 704R	Operational amplifier	18	18	68
Motorola MC 1530	Operational amplifier	280	4400	1700
T. I. 709L	Operational amplifier	800	3600	2680
Radiation, Inc. RD211†	Dual quad-diode gate expander	70	20	--
Radiation, Inc. RD220†	Hex inverter (digital)	36	138	380
Radiation, Inc. RD221†	Dual binary gate	270	180	490
Radiation, Inc. RA239†	Amplifier (analog)	—	80	66
Philbrick Q25AH†	Hybrid amplifier	200	16	320
Philbrick Q25AH†	Hybrid amplifier	100	2000	1000
Fairchild μA709	Operational amplifier	11	30	—

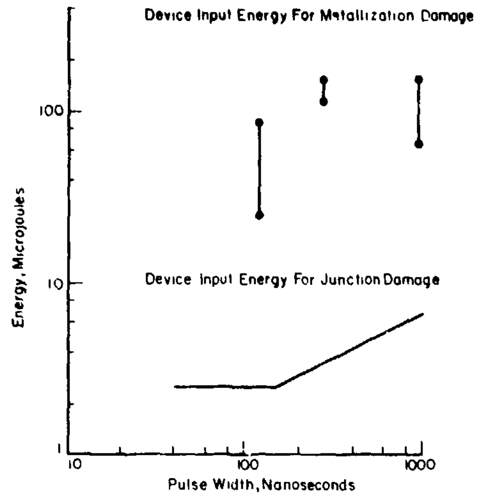
*Applied pulse of 1-μs duration

†Normalized to 1-μs pulse (original data for 0.1 μs)

Interconnection Failure Modes

The vulnerability of device leads, metalization patterns and lead bonds, for the most part, can be considered as a thermal problem. In this case, the problem reduces to that of considering heat dissipation due to system thermal conductivity up to its melting point, together with an assessment of the dynamic stress conditions produced at material discontinuities. In general, one would expect that at least the leads and metalization patterns should exhibit a fairly uniform current density throughout their material cross sections for relatively moderate pulse widths as compared to the current constriction sites in semiconductor junctions which can be altered by defect and bias conditions. For relatively short pulses, such a phenomenon as "skin effect" would, of course, alter the cross sectional current density in such a way as to produce a "peripheral current constriction" condition. These effects, though, are fairly well defined and can be considered in a rather straightforward manner. The lead bonds and any multi-metal metalization patterns can also be considered in somewhat the same fashion. However, bond interface impedance, possible current constriction sites, and hydrodynamic pressure pulses due to interface discontinuities may also have to be considered. In general, it is observed that the vulnerability of the interconnection system usually occurs at current levels in excess of those required to cause significant junction damage in typical semiconductor devices at hundred nanosecond pulse widths.

Typical EMP pulse type experimental data is shown in the following figure. This type of data is in direct contrast with data obtained when a component is exposed to high frequency RF pulses as in the case of testing done to ascertain the hazards to ordnance from electromagnetic radiation (HERO). In this case, much of the energy is shunted around the junction due to its capacitance, and the incipient degradation in most cases is due to lead melting.



EXPERIMENTAL INPUT ENERGY FOR JUNCTION AND METALLIZATION DAMAGE - RO 211 MICROCIRCUIT

An analytical model and the use of the model are presented in other references.

Synergistic Effects

An additional area of concern in the case of semiconductor failure has been that of synergism between the electrical overstress due to an EMP and the gamma ionizing dose rate ($\dot{\gamma}$). While exhaustive studies of synergism have not been conducted, four independent and reliable investigations have shown no evidence, or only weak evidence, of synergistic effects in discrete semiconductors or IC's.

In the case of complete systems, however, the $\dot{\gamma}$ triggering a circuit into conduction can permit damage due to electrical pulses.

SYNERGISM

COMPONENT LEVEL

THREE INDEPENDENT AND RELIABLE INVESTIGATORS HAVE SEEN NO EVIDENCE OF SYNERGISM BETWEEN ELECTRICAL OVERSTRESS PULSES AND $\dot{\gamma}$

VAULT (HDC) - INVESTIGATIONS ON IC'S SHOWED NO MORE THAN A FACTOR OF 2 DIFFERENCE IN INDIVIDUAL ENVIRONMENT FAILURE LEVELS AND COMBINED ENVIRONMENT FAILURE LEVELS

BUDENSTEIN (AUBURN) - INVESTIGATIONS ON SOS DIODES SHOW $\dot{\gamma}$ COOLS A JUNCTION APPROACHING HOT SPOT NUCLEATION

RAYMOND (NORTHROP-MRC) - EXTREMELY WEAK EVIDENCE OF SYNERGISM IN IC'S

HABING (SANDIA LABS) - REALISTIC ENVIRONMENT TEST SHOWED NO SYNERGISM EFFECT WITHIN A FACTOR OF 2 OR LESS

SYSTEM LEVEL

IONIZING RADIATION CAN TRIGGER CONDUCTION WHICH PERMITS ELECTRICAL PULSES TO CAUSE BURNOUT

SUBJECT TO DESIGN RULE DOCUMENTATION

Resistor Failure

Resistive elements in the form of either lumped resistors or diffused resistors often are the terminating element for long cables. Consequently, information on the way these devices fail and typical failure levels are important. Tests have been conducted on wire wound, metal film, carbon composition and diffused resistors. The failure levels vary with number of pulses, duty cycle, and power or voltage. Conventional ratings indicate that for pulse applications, the pulse power rating is 10 times the dc power rating and the voltage rating is 1000 volts per inch. For low duty cycle, as in the case of EMP, these are far too conservative.

Failure Modes

Four types of failure have been found. These are:

1. Resistance value change - failure is defined as a change in value beyond normal tolerance. The importance of this change is dependent on the circuit function. This mode of failure can be due to thermal effects (energy dissipation) or voltage stress induced.

2. Internal breakdown - this break down occurred when the resistor under test opened but did not blow apart or no external evidence of arcing was present. This was due to thermal dissipation with the device.

3. Arc across resistor casing - this type of breakdown was exemplified by an arc across the external surface of the resistor. No damage to the resistor resulted from this failure.

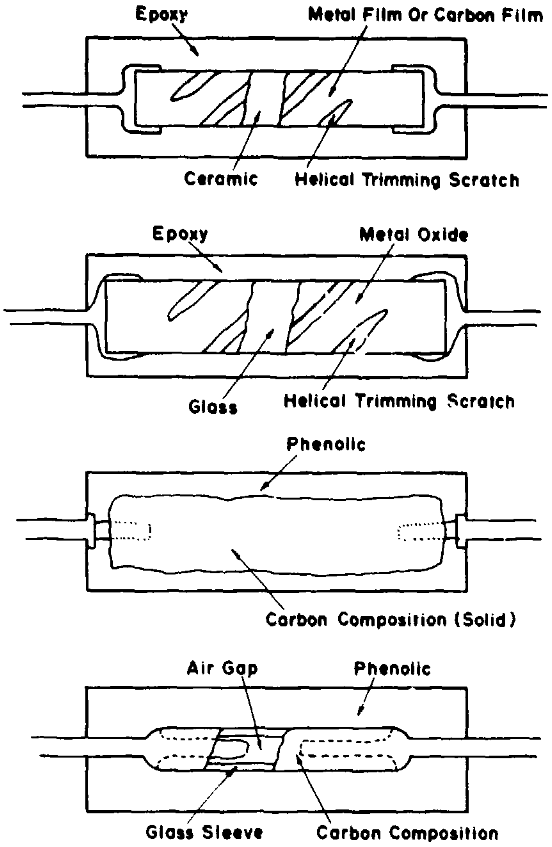
4. Catastrophic breakdown - this type of breakdown occurs when an external arc starts across the resistor, but due to some defect in the ceramic casing, re-enters the core. The pulse energy is then dissipated in only a small fraction of the resistor and causes the casing to rupture (blow off) and the resistor to open.

All of these failure modes have been seen in the resistor tests. To define a safe working voltage level for the resistor, it was given as the level where the resistance did not change as a result of the applied pulses. These data are reported in the next paragraphs.

Resistor Construction

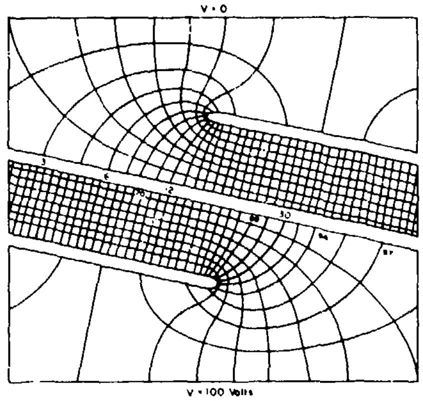
The construction technique plays an important role in the failure mechanisms or level for resistors. Typical resistor constructions are shown in the following figure.

CROSS SECTION OF RESISTOR CLASSES



Metal film, carbon film, and metal oxide resistors are constructed with the film deposited, or the oxide grown on a glass or ceramic substrate. The thickness of the film or oxide layer determine the minimum resistance value of the resistor. To increase the resistance, spiral cuts are made in the film to increase the total path length of the current. A very limited family of film thickness is used to cover many decades of resistance value. Spiraling of the film results in uneven voltage distribution across the resistor body which results in voltage breakdown at lower levels than would normally be expected. This breakdown normally occurs at the ends of the spiral where the voltage gradients are highest. The voltage distribution for a typical spiraled film resistor is shown in the following figure.

EQUIPOTENTIAL LINES AND FLOW LINES IN THE RESISTIVE LAYER OF A SPIRALLED FILM RESISTOR. THE NUMBERS REFER TO POTENTIALS.(REF 3:14)



In the case of carbon composition resistors, the entire body of the device is the resistance material, the conductivity of which determines the resistance value and the volume of the power dissipation. The failure level in this type is strongly influenced by the geometry of the lead connection to the body since this determines the current distribution and voltage gradient at the interconnection. Large contact area, that is, expanding the lead at the connection point, results in the maximum useful carbon volume and the minimum voltage gradient.

The composition resistor shown at the bottom of the previous figure utilizes a carbon composition material on a hollow glass substrate. The terminating ends of the lead connections are in the resulting air gap. These type resistors exhibit voltage breakdown in the air gap with no ultimate effects on the resistor.

Resistor Failure Threshold

The susceptibility of resistors can be categorized by resistor type. The data are very consistent for the same manufacturer. Variations between manufacturers is usually due to the geometry of lead connections and construction variations. The susceptibility ranking for metal film and metal oxide resistors is almost equal.

RESISTOR SUSCEPTIBILITY

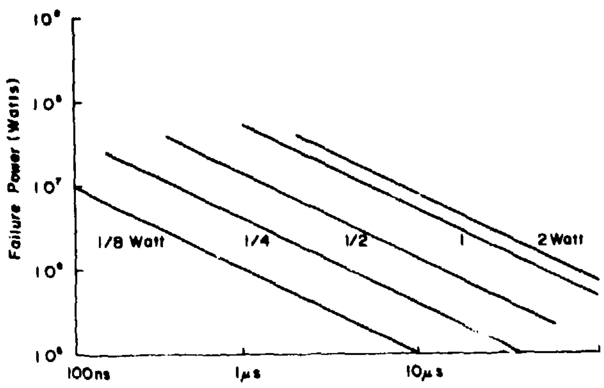
- **Wire Wound---------- Hardest**
- **Carbon Composition**
- **Carbon Film**
- **Metal Film**
- **Metal Oxide---------- Softest**

The resistor failure data for all types, with the exception of the wire wound resistors, indicate energy dependence (i.e., adiabatic heating, $t^{-\frac{1}{2}}$) for pulse widths between approximately 100 nanoseconds and 50 microseconds (or greater). This range of pulse widths covers the EMP induced transients anticipated, so for EMP damage thermal failure is usually the normal mode.

Carbon Composition Resistors

For carbon composition resistors, the failure power is proportional to the device rated power. For low wattage rated resistors (< 1 watt) the failure power is directly proportional. Above 1 watt, the failure power does not double as the rated power is doubled. This is due to some nonuseful carbon volume for pulsed signals due to lead connection techniques. The failure mode for thermal failure is a melt (melt fingers being formed) at the boundary between the interconnecting wire and the bulk material. At shorter pulse widths (less than 100 ns) the failure mode is voltage dependent in that surface or internal arcing occurs. This is usually seen as a sharp fracture in the bulk material.

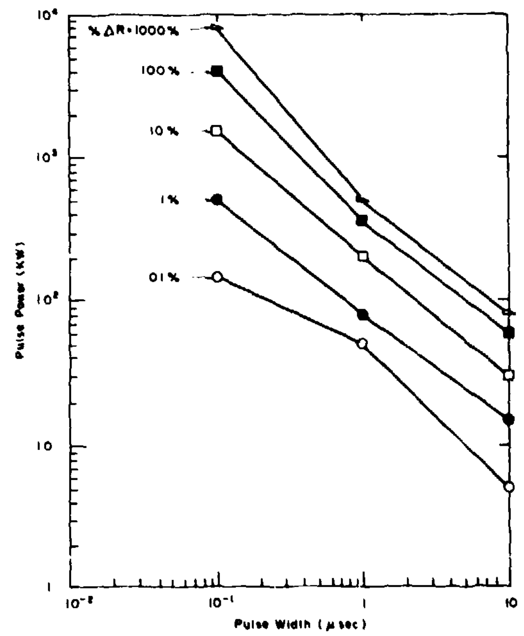
CARBON COMPOSITION RESISTOR FAILURE MODELING



Film Resistors

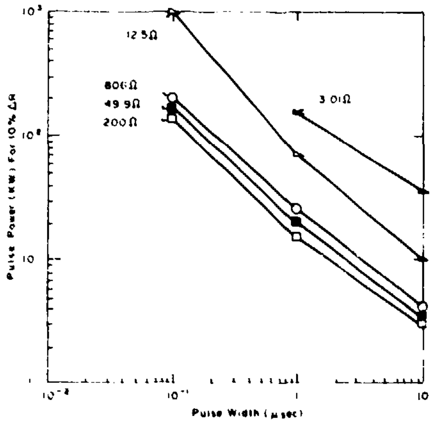
Film resistor failure breaks from the adiabatic curve ($t^{-\frac{1}{2}}$) at pulse widths between 10 and 50 microseconds. The coating type (metal, metal oxide or carbon) and technique change the thermal properties of the film and, consequently, its failure level. For damage defined as a change in initial resistance, carbon film resistor damage (%ΔR) increases monotonically with pulse power and pulse width (energy). For a given damage level (%ΔR) the pulse power for failure decreases monotonically with pulse width.

RESISTOR DAMAGE EXTENT AS A FUNCTION OF DAMAGE POWER AND PULSE WIDTH FOR DALE MC 1/4 (0.25 WATTS) 49.9 OHM CARBON FILM RESISTORS



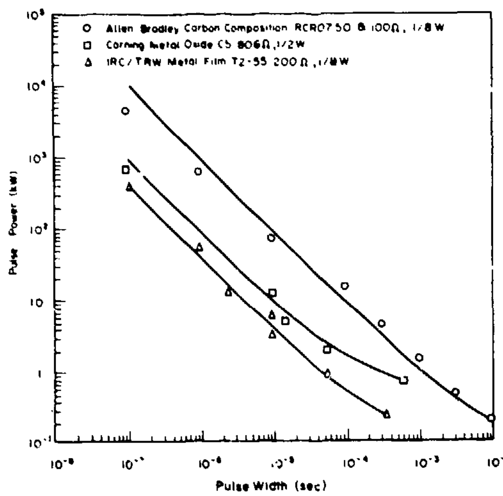
The pulse power required for a given damage level (10% Δ R) is also a function of the initial resistance of the resistor. The damage power decreases monotonically as resistance value increases from 3 ohms to 200 ohms. It then increases for an initial resistance value of 806 ohms. The 3 to 200 ohms resistors had the same film thickness, the resistance value being determined by the amount of spiraling. The 806 ohm resistor was a different film thickness (thinner to increase the resistance) with less spiraling to achieve the final resistance value. It is apparent that the greater the amount of spiraling required to achieve the final resistance value, the lower the failure threshold. This is because of the greater current concentrations due to the spiraling.

DAMAGE POWER DEPENDENCE ON PULSE WIDTH AND INITIAL RESISTANCE VALUE FOR DALE MC 1/10 (0.1 WATTS) CARBON FILM RESISTORS



The metal film and metal oxide resistors exhibit lower failure thresholds than the carbon film or composition resistors. The failure power for metal film, metal oxide, and carbon composition resistors is shown in the following figure.

DAMAGE POWER DEPENDENCE ON PULSE WIDTH

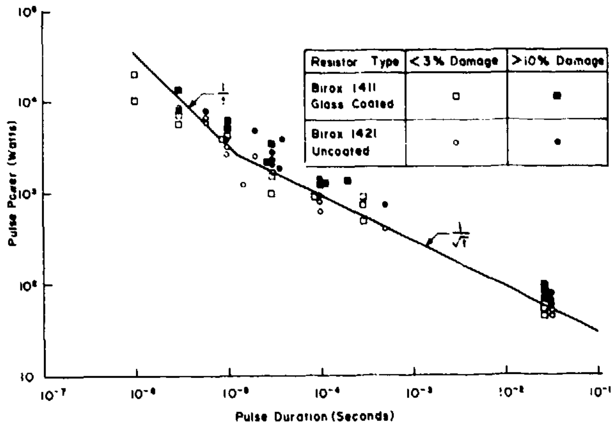


The failure threshold of the metal oxide resistor appears to be higher than for the metal film resistor. Note that the metal film and carbon composition resistors are both 1/8 watt rating while the metal oxide has a 1/2 watt rating.

Diffused Resistors

The failure thresholds for diffused resistors is of the same order of magnitude as for metal film and metal oxide resistors. The failure threshold is energy dependent (adiabatic heating) following a $t^{-\frac{1}{2}}$ slope for pulse widths between 10 microseconds and 0.1 seconds. For pulse widths below 10 microseconds, the failure threshold curve follows a t^{-1} slope.

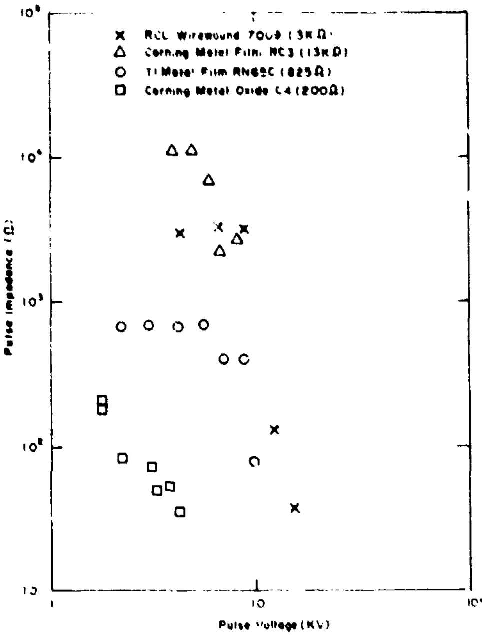
THERMAL FAILURE IN DIFFUSED RESISTORS



Short Pulse Width Failure

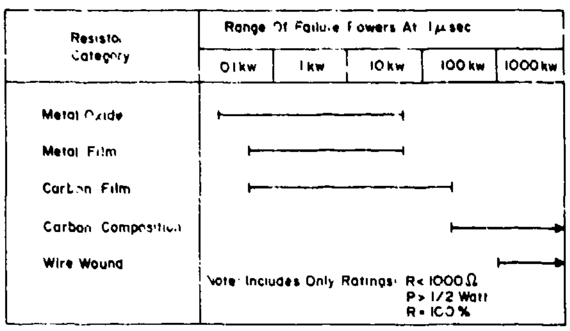
For pulse widths less than approximately 100 nanoseconds, the failure (Δ R) is voltage dependent. This results from the extremely high voltages required to deliver sufficient energy for thermal failure for short pulse widths. This type of failure may be seen in all types of resistors, but is most prevalent in wire wound and film resistors. In carbon composition resistors, the failure is usually surface breakdown. In wire wound or film types, the arcing occurs between turns or across the boundary between spirals. It is manifest in the form of an immediate reduction in the resistor pulse impedance as shown.

VOLTAGE LEVEL EFFECTS ON RESISTOR PULSE IMPEDANCE



Summary of Resistor Failure Thresholds

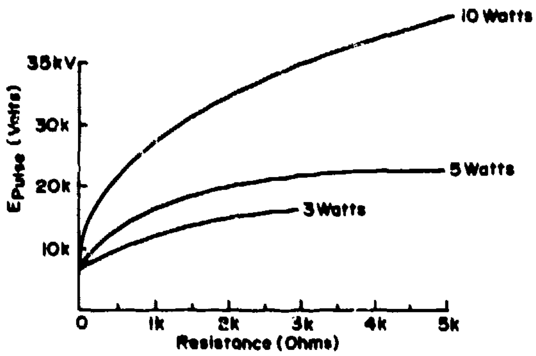
A summary of measured resistor failure thresholds is presented in the following figure. Since the data were obtained using 1 microsecond pulses, the power indicated for failure can be equated to a damage constant (K) as for semiconductors. An assigned K value for this pulse width would be numerically equal to the power in kilowatts (i.e., for a threshold of 1 kW, k = 1).



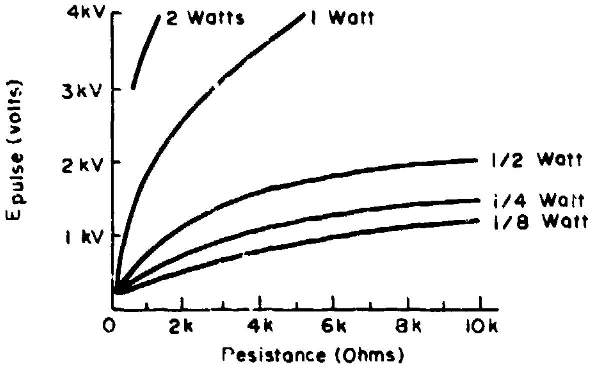
RANGE OF RESISTOR FAILURE POWER

Resistor Failure Thresholds Based on Safe Operating Voltage

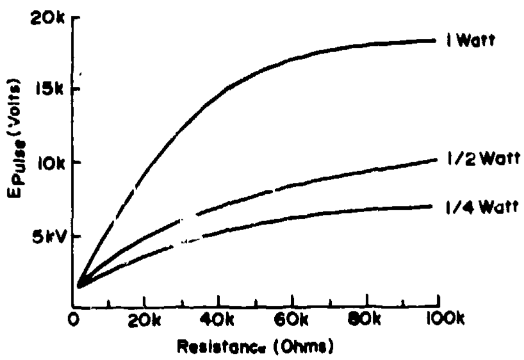
The maximum safe pulse voltage for various types of resistors as a function of the resistance value, with the wattage rating as a parameter are presented in the following figures. The average pulse power in no case exceeded the dc power rating for the resistor. The pulse duration used for obtaining these data was 20 μsec.



WIRE-WOUND RESISTORS



METAL FILM RESISTORS

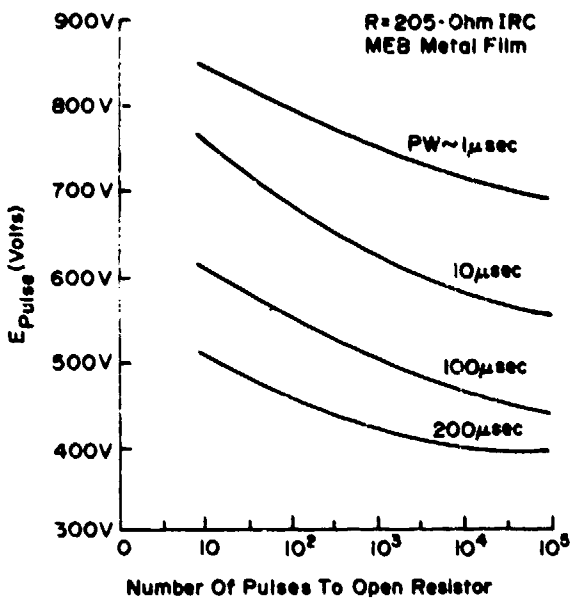


CARBON-COMPOSITION RESISTORS

The pulse power rating is approximately 5000 times the dc rating for wire wound resistors, 1000 times the dc rating for metal film resistors and 500 times the dc rating for carbon composition resistors. The pulse power rating for these devices can be determined approximately from

$$P_p = \frac{E_p^{\,2}\ \text{(volts)}}{R\ (\Omega)}\ \text{watts}$$

It should be noted that for carbon composition and metal film resistors of low resistance value and low wattage rating, failure can occur at voltages of a few hundred volts on a single pulse basis. This failure level is further reduced for multiple pulses or increased pulse duration.



PULSE WIDTH VS MAXIMUM PULSE VOLTAGE

Capacitor Failure

Tests on capacitors have been limited in types of studies and component sample size because they have been considered to be much harder than other electronic components such as semiconductors and thin film resistors. These limited studies have indicated that some capacitor types fail at levels as low as those seen for semiconductors. Therefore, consideration should be given to the failure levels of these components because if the semiconductors in the circuit are protected, the nonsemiconductor components may determine the resulting EMP vulnerability.

Failure Modes

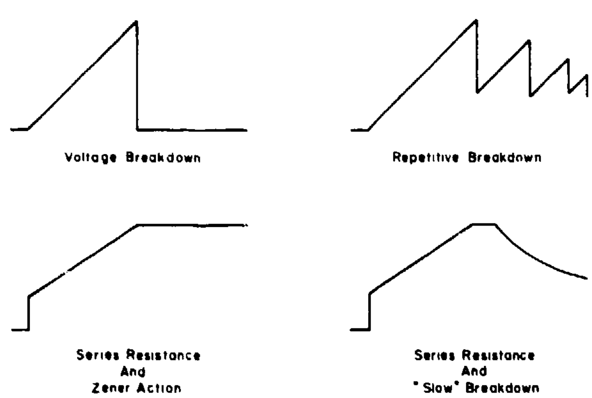
The basic failure mechanism in capacitors is internal (dielectric) breakdown. The parameter that has been found to change is the dissipation factor (D) due to a change in the leakage resistance (R). This relationship is given by

$$D = \frac{\omega C}{R}$$

For ceramic type capacitors this breakdown is very abrupt. The amount of post breakdown degradation was related to the energy dissipated in the capacitor during breakdown. The breakdown voltage was lowered after the initial breakdown occurred (i.e., for subsequent pulses). In some cases (manufacturers' types), no evidence of changes in the capacitor parameter was seen; whereas in other cases, the dissipation factor increased by a factor of 100.

The upper curve shows the characteristic breakdown of most types of capacitors. Once breakdown occurs (in capacitors such as paper or disc ceramics) it is usually sustained. The post breakdown degradation (that is any self healing ability, decrease in breakdown voltage, etc.) depends on the total breakdown energy. The exciting pulse was a double exponential waveform.

BASIC RESPONSE CHARACTERISTICS EXHIBITED BY CAPACITORS FOR SQUARE WAVE CURRENT PULSES

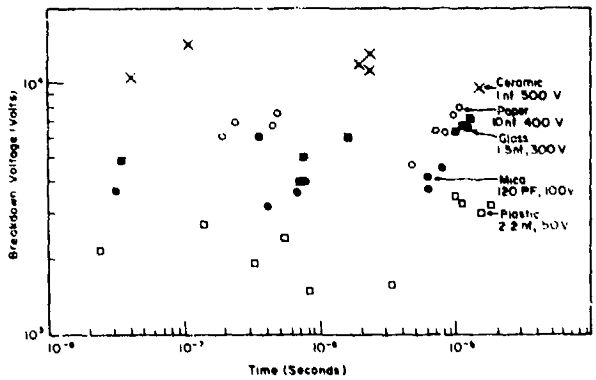


In the case of low voltage tantalum electrolytics, the breakdown characteristics are quite different. The breakdown in this case was a slower process as shown in the lower curves. The abrupt breakdown was not observed but the leakage resistance decreased progressively until breakdown occurred. As the leakage current increases, dissipation in the device increases, the sustaining voltage decreases. In other tests, abrupt changes were seen but, in all cases, were preceded by large leakage currents. After breakdown shots, dissipation factors often rose by as much as a factor of 600, with the equivalent resistance dropping as low as 1.5 ohms.

Failure Thresholds

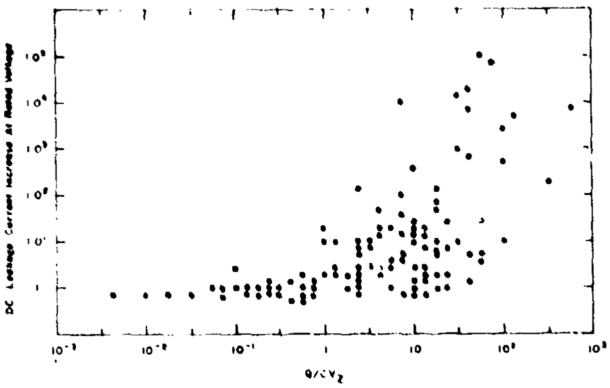
The voltage failure thresholds versus pulse rise time for ceramic, paper, glass, mica, and plastic are presented in the following figure. It is apparent from the data that their is only a slight voltage turn up for short (< 100 ns). The failure thresholds are many times (15 to 40) the rated working voltage of the capacitors and, therefore, are relatively hard to EMP induced transients.

PULSE RISE TIME EFFECTS ON VOLTAGE BREAKDOWN IN



The damage to tantulum capacitors is generally seen as an increase in leakage current. As was seen in the breakdown waveforms, tantulum capacitors have a zener voltage (Vz) characteristic which is equal to the forming voltage (voltage at which oxide was formed). Breakdown occurs slowly (microseconds) when voltage exceeds the forming voltage. The increase in leakage current is a direct function of the total charge transferred during the incident pulse. The increase in leakage current at rated voltage as a function of total charge transferred is shown in the following figure.

LEAKAGE CURRENT DAMAGE IN TANTALUM FOIL ELECTROLYTIC CAPACITORS POSITIVE POLARITY PULSING



The forming voltage is always higher than the rated voltage. The forming voltage, however, is not a fixed percentage. For higher voltage units, it is a much smaller percentage than for low voltage units and, therefore, has a smaller safety margin. The forming voltage is not normally indicated on data sheets, but may be obtainable from the manufacturers. For the tantulum capacitors tested, the minimum breakdown voltage was approximately two (2) to three (3) times the rated voltage as indicated in the table.

SOLID TANTULUM ELECTROLYTIC CAPACITOR SPECIFICATIONS, TEST CONDITIONS, AND DETERMINED BREAKDOWN VOLTAGES

Sprague Device No	Polarity	Capacitance (μF)	Voltage (WVdc)	Breakdown Voltage Mean (V)	Breakdown Voltage Standard Deviation (V)	Breakdown Voltage Minimum (V)	Pulse Width (μs)	Tested Devices (No)
472X9035A2	Forward	0.0047	35	155.0	43.1	90.0	2.6/4	19
225X9035B2	Forward	2.2	35				4.8/30	24
225X9015A2	Forward	2.2	15	143.0	48.7	68.0	3/30	17
472X9035A2	Inverse	0.0047	35	106.0	19.7	65.0	1/10	15
225X9015B2	Reverse	2.2	35				3/30	15
225X9015A2	Reverse	2.2	15	53.7	6.5	43.0	30	6

Low voltage tantulum capacitors can fail at energy levels comparable to those of semiconductor devices. The minimum energy levels at which failure was seen for tantulum capacitors is shown along with typical failure energy levels for semiconductors in the figure.

TYPICAL ENERGY FAILURE LEVELS OF SEMICONDUCTORS COMPARED TO THE ENERGY REQUIRED TO DAMAGE LOW-VOLTAGE TANTALUM CAPACITORS

Component	Energy (μJ)
Point Contact Diode IN82A-2N69A	0.7 to 12
Integrated Circuit A709	10
Low-Power Transistor 2N930-2N1116A	20 to 1000
High-Power Transistor 2N1039 (Ger)	1000 and Up
Switching Diode IN914-IN933J	70 to 100
Zener Diode IN702A	1000 and Up
Rectifier IN537	500
Solid Tantalum Capacitor	61 and Up

The Semiconductor Data Were Based On A 1μs Damaging Pulse (From D Tasca, Document No 70SD401, The General Electric Company (January 1970) And Unpublished Data By J R Miletta)

Inductive Elements

Inductive elements can fail in the way similar to capacitors and resistors whereby a temporary impairment such as an arc-over or saturation occurs such that the characteristics of the inductive elements are not impaired on a long-term basis. Similarly, a catastrophic failure can occur such as an arc-over and punch-through for the insulation similar to the capacitor insulation or semiconductor surface failures.

Studies to date on inductive elements per se have been quite limited owing to the relative hardness of these devices in comparison to the more susceptible semiconductors and passive thin film resistor elements.

Squibs and Detonators

Squibs and detonators can play an important role in the EMP susceptibility or vulnerability of a particular system. Typically, squibs and detonators are not only used to initiate the formation of a final explosive, but also to perform separation of stages in the initiation of rockets. In general, squibs and detonators can fail in two ways. First of all is the premature unwanted ignition of the pyrotechnical devices with obvious catastrophic implications. The other is dudding of the device such that after EMP exposure, the device no longer performs in a satisfactory manner.

Major emphasis in the past has been directed toward protecting the squibs and detonators from either dudding or unwanted pre-ignition from either static charges or energy picked up under microwave illumination conditions. The EMP failures of the device can occur when sufficient energy from EMP pickup is applied to one or both terminals of the pyrotechnical device. Typically, the two terminals of the device are fired via a balanced two-wire cabling system. Here, the differential mode pickup under EMP conditions may be somewhat lower than the common mode pickup. In the case of common mode pickup, one observed failure mode occurs when the arrangement is such that an arc-over can occur between one of the wires to ground and not the other. This, in effect, converts the common mode into a differential mode of sufficient magnitude to cause premature detonation. In other cases, arc-overs can occur within the case between the sensitive elements of the pyrotechnical device and the case.

EED's (electroexplosive devices) are the more sensitive to ignition requiring for the most sensitive devices only a few ergs. EBW's (exploding bridgewires) require considerably more energy, minimum energy values being on the order of five (5) millijoules. Bridgewire burnout always results in ignition of EED's but in the case of EBW's may only result in dudding the device.

Terminal Protective Devices

Terminal protective devices can also be damaged by electrical overstress or excess energy during conduction. These devices include zener type devices, gas tubes, spark gaps, thyristors, etc. As seen in the following table, the voltage failure level and associated pulse durations for typical TPD's are higher than the normally anticipated EMP induced transients. Failure of these devices, however, could possibly occur for very long transmission lines, large antennas, etc.

APPROXIMATE FAILURE LEVELS

DEVICE	V_B(V)	FAILURE LEVEL (V)	PULSE DURATION µS
LOW VOLTAGE ZENER	6.8	11,000	50/500
" " "	20	11,000	50/500
" " "	200	11,000/11,000	50/500
BIASED ZENER LIKE	20	3800	50
" " "	200	3800	50
GAS TUBE	300-500	3800	50
" "	500-900	3800	50
SPARK GAP	470	11,000	50/500
" "	2500	11,000	50/500
THYRISTOR	60	3800	50
"	50A	3800	50

Miscellaneous Devices

A comparison of the test results for a variety of devices is shown in the following table. As can be seen from these data, most of these devices are hard to the the normally anticipated EMP induced transients, although the potential for failure cannot be excluded in all cases.

These data are a summary of very limited test results since most of these devices have been considered to be inherently hard.

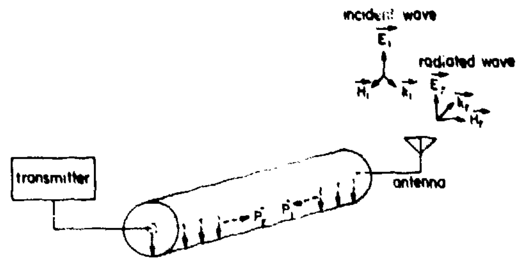
TEST RESULTS FOR MISCELLANEOUS COMPONENTS

Component Type	Test Results	Comments
1 Spark Gaps - Rated 75V to 5kV	No failure 1500 amps, 160 joules	Hard to essentially all EMP transients
2 Magnetic Surge Arrestors - Rated 500 V to 8kV	No failure to 1000 amps, 160 joules	Hard to essentially all EMP transients
3 Neon Lamps Rated 55V to 110 V	No failure to 1000 amps, 160 joules	Hard to essentially all EMP transients
4 Varistors		
• MOV	No failure to 1000 amps, 160 joules, clamping voltage increase 1600 amps, 50 joules	Hard to essentially all EMP transients
• Thyrite	Failure at 300 amps, 22 joules 10 amps, 50 joules	Possibly susceptible to very high signal levels
5 Relays, Motors, Transformers, Switches, and Potentiometers	No failure at 1kV 0 8 joule	Hard to at least 1 kV
6 Discrete Filters	Failure at 5kV, 0 8 joule	Hard to 5kV
7 Filter Pin Connectors	Leakage increases at 300V, 0 04 joule	Potentially susceptible at moderate levels (>300V)
8 Tubes	Degradation at 1kV, 0 8 joule	Potentially susceptible at levels > 1 kV
9 EED's	Activation at 2kV, 6 joules	Susceptible at levels > 2kV

5.4 CABLE AND CONNECTOR FAILURE

Other very important system components that may be functionally damaged by EMP induced transients are cables and connectors. This is particularly important for cables that are already electrically stressed such as transmitter output cables. The difference between the voltage applied to the cable by the transmitter and the voltage breakdown rating of the cable may be sufficiently small that EMP induced transients will cause breakdown of the insulation or air space in connectors.

TRANSMITTING SYSTEM CABLE WITH VOLTAGE APPLIED BY THE TRANSMITTER AND ANTENNA

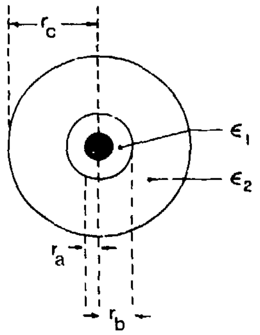


field due to transmitter applied voltage plus antenna applied voltage

The breakdown strength of cable insulation may be limited by the dielectric strength of small imperfections in the insulation. Within the body of the cable, breakdown starts from a small air pocket. At first, discharges take place in the pocket. This produces local heating. The insulation melts and carbonizes and ultimate failure occurs, either through mechanical effects or due to a short-circuit produced by a carbonized track across the dielectric from one conductor to another. A mechanical effect that can occur is for distortion of the dielectric to occur resulting in the inner conductor becoming eccentric and touching the outer conductor.

The role of dielectric imperfection in producing cable breakdown is shown in the figure which portrays the cross sectional view of a coaxial cable at a site where an imperfection in the cable dielectric is assumed to exist Here it is assumed that the insulator with a dielectric constant of ϵ_2 does not touch the center conductor due to a manufacturing defect. Hence, an air pocket with a dielectric constant of $\epsilon_1<\epsilon_2$ exists around the center conductor. Due to the dissimilar dielectric constants, the electric field in the air pocket will be enhanced, thus causing an initial arc in the air and a subsequent breakdown in the insulating material.

ELECTRIC FIELD DUE TO TWO DIELECTRIC MATERIALS IN CONCENTRIC COAXIAL CABLE



Cable connectors, in many instances, breakdown prior to cable failure. This is because of the inadvertent air paths that exist in many connectors due to the construction techniques. As in the case of cable failure, it is these air paths that result in breakdown.

5.5 OPERATIONAL UPSET MECHANISMS

As mentioned previously, operational upset is primarily a circuit or system problem. In general, it is not related to individual components comprising the circuit but rather depends on the circuit function, circuit operating levels (biases), the circuit type (digital or analog), and the nature of the waveform driving the circuit.

Operational upset can occur in both digital and analog circuits. In analog circuits an EMP transient may be amplified and interpreted as a control signal, or it may be interpreted as a fault current resulting in circuit breaker operation, or result in the opening of fuses if the currents persist for a long enough period and contain sufficient energy. Low level pulses in analog circuits usually appears as noise and does not interrupt circuit operation. In digital circuits, the induced waveform may be interpreted as discrete pulses which are propagated through the system resulting in errors. For example, flip-flops and Schmitt triggers may be inadvertently triggered, counters may record wrong counts, or memories may be altered due to driving current or direct magnetic field effects. These voltage and current thresholds are usually much lower than those required for analog circuit upset so digital circuits and semiconductor or core memories are the most susceptible.

Digital Circuit Upset

Operational upset mechanisms will be briefly illustrated by considering the effects of EMP-induced transients on digital logic circuits and computer memories. A digital logic circuit may be upset by input terminal disturbances or by dc and ground disturbances. The problem is further complicated by whether or not the disturbance propagates through the system.

DIGITAL LOGIC CIRCUIT UPSET

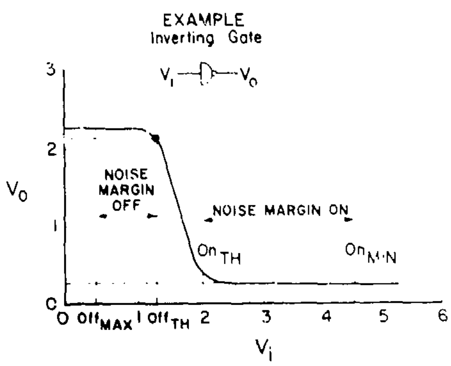
There are two general types of problems associated with upsetting a digital logic circuit:

1. input terminal disturbances
2. DC power and ground disturbances

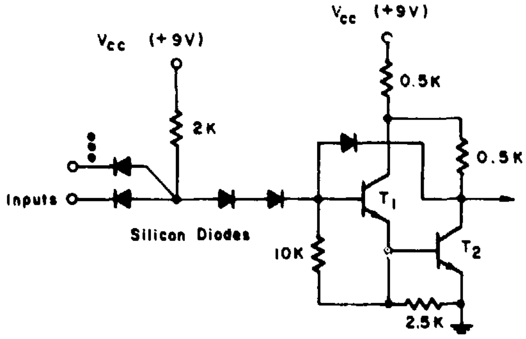
An example of operational upset is given by considering an inverting gate. An unwanted pulse on such an inverting gate may change its state. This undesired change of state of the output of the inverting gate may be amplified by the following gate and propagated on through a string of digital gates. An error may thus arise in a bit in a data register.

An input disturbance will be propagated by the following gate if the undesired output state of the first gate exceeds the on-threshold of the second gate.

DC TRANSFER FUNCTION OF AN INVERTING GATE

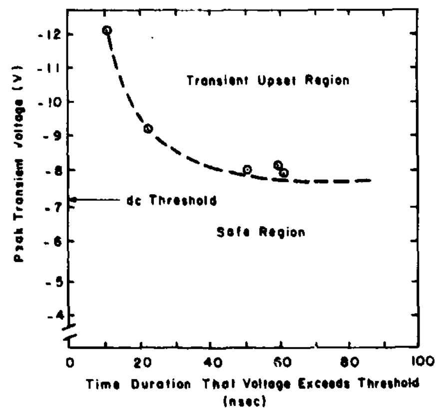


The circuit diagram for a nonsaturating DTL (diode-transistor logic) dual four-input gate, constructed with dielectric isolation and thin film resistors, is shown in the accompanying figure. It has a low noise margin (1.2 v) and a high speed (10 nsec) propagation time.

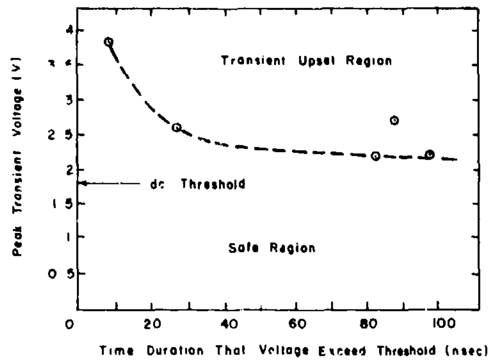


DTL GATE CIRCUIT

If the semiconductor forward voltage drops are all equal to $V_\phi \approx 0.6$ volt, the input dc threshold for conduction of T_1 and T_2 is approximately $3V_\phi = 1.8$v. Conversely, the (largest) threshold for turning T_1 and T_2 off is $3V_\phi - 9 = -7.2$ v This means that the transient necessary for turn on must be at least 1.8 v, and for turn off at least -7.2 v. Unbalance of this sort is undesirable from an EMP hardness standpoint. For transients that exceed the threshold for times less than the specified propagation delay, a higher level can be tolerated before transient upset occurs. Although this level is related to the time duration, the polarity, the input point, and circuit parameters such as noise immunity and response speed, no straightforward way of establishing the relationship is known except to experimentally test the circuit. Experimental results for this circuit for negative and positive upsets on the input lead are shown in the following figures. Once the transient pulse width exceeds the propagation time, the required upset voltage asymptotically approaches the dc threshold voltage. Shorter pulse widths require correspondingly greater voltages.



INPUT LEAD NEGATIVE UPSET



INPUT LEAD POSITIVE UPSET

These results show that the dc threshold establishes a safe measure for transient upset. Observations indicate that transient upset levels appear to be independent of the exact waveshape, depending rather on the peak value. It has also been observed that circuit threshold regions for upset are very narrow. That is, there is a very small amount of voltage amplitude difference between the largest signals which have no probability of causing upset and the smallest signals which will certainly cause upset.

Memory Erasure

Computer memories are also susceptible to EMP induced transients. The level of susceptibility is determined by the magnetic field required to change the magnetization state of the memory element (magnetic memories), or the voltage or current thresholds for semiconductor memories.

In the case of magnetic memories, the magnetic field impressed may be due to direct magnetic field effect (impinging magnetic field illumination) or driving currents induced in the associated wiring. The direct effects require much higher levels of incident magnetic field than the induced current effects. The table indicates the minimum energy levels required for upset of typical digital circuits and memories due to EMP induced signals.

The minimum energy necessary for operational upset is on the order of one to two orders of magnitude less than for damage of the most sensitive semiconductor components.

MINIMUM ENERGY TO CAUSE CIRCUIT UPSET OR INTERFERENCE

DESIGNATION	MINIMUM ENERGY JOULES	MALFUNCTION	OTHER DATA
Logic Card	3×10^{-9}	Change of State	Typical logic transistor inverter gate
Integrated Circuit	4×10^{-10}	Change of State	Sylvania J-K flip-flop monolithic integrated circuit (SF 50)
Memory Core	5×10^{-8}	Core Erasure Via Wiring	Burroughs medium speed computer core memory (FC 8001)
Memory Core	3×10^{-9}	Core Erasure Via Wiring	RCA medium speed, core memory (269M1)
Amplifier	4×10^{-21}	Interference	Minimum observable energy in a typical high gain amplifier

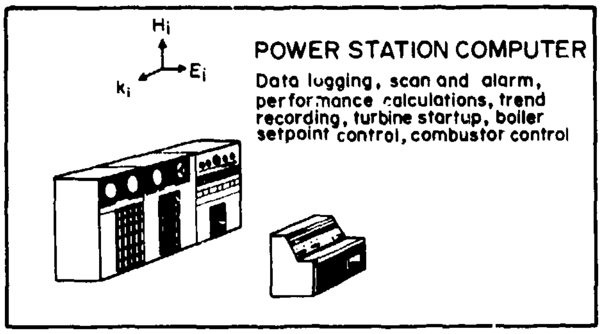
The most susceptible memories are those that require the smallest driving currents to cause a change of state in the memory, such as core or semiconductor memories. It should be noted, however, that if the transients are induced in circuits external to the memory proper, that is, prior to the read/write amplifiers for example, they may be amplified in the same manner as normal signals and written into memory. Under this condition, the hardness of the memory proper is a secondary consideration. It is obvious that under this condition the memory is most susceptible in the write mode of operation. This has been verified experimentally.

Effects of Operational Upset

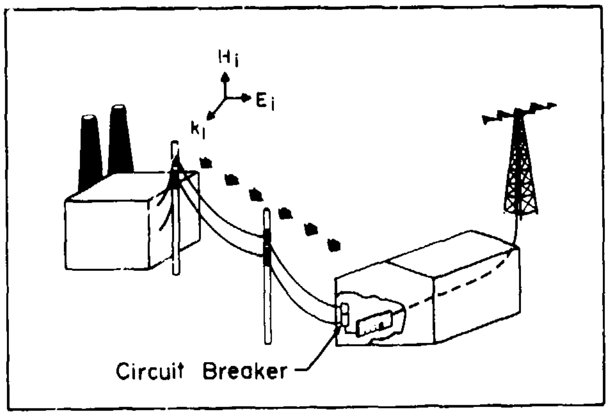
The effect of operational upset on system performance and mission is highly dependent on the system design and use. In some systems, loss of synchronization for as long as a few milliseconds is of no great importance. On the other hand, loss of stored information in a computer may require restart of a very long computer program, thus delaying the operation of a specific system.

The trajectory control of a spacecraft or missile is an example where operational upset for a very short time may be of considerable significance.

Some functions performed by a computer may be unimpaired by relatively long periods of circuit upset. Others may be impaired by short periods of upset. In this example of a power station computer, functions such as data logging, scan and alarm, performance calculations, and trend recording may be relatively unaffected by operational upset. On the other hand, process control functions such as turbine startup, boiler setpoint control, and combustor control may be affected to a much greater extent.



Large amounts of energy may be collected by power lines and cause circuit breakers to open. The time to re-energize the system may cause its function to be seriously impaired. Also, since a considerable amount of a generator's load could be dropped, undesirable effects might occur in the generating and transmission system.



REFERENCES

Burdenstein, Paul P., *et al.*, "Second Breakdown and Damage in Semiconductor Junction Devices," Auburn University, Auburn, Alabama, Report No. RG-TR-72-15, April 1972.

"EMP Engineering and Design Principles," Bell Telephone Laboratories, Loop Transmission Division, Electrical Protection Department, Whippany, New Jersey, 1975.

Lennox, C.R., "Experimental Results of Testing Resistors Under Pulse Conditions," Electrical Standards Division 2412, Sandia Laboratory, Albuquerque, New Mexico (PEM #6).

Case, C., Miletta, J., "Capacitor Failure Due to High-Level Electrical Transients," HDL-TM-75-25, Harry Diamond Laboratories, Adelphi, Maryland, December 1975.

"DNA EMP Preferred Test Procedures," IIT Research Institute, DNA 3286H, Chapter 14 (to be published).

"Electromagnetic Pulse Handbook for Missiles and Aircraft in Flight," Sandia Laboratories, AFWL-TR-73-68, EMP Interaction Note 1-1, September 1972.

Domingos, H., "Pulse Power Effects in Discrete Resistors," Clarkson College of Technology, Potsdam, N.Y., AFWL-TR-76-120, November 1976.

Tasca, D.M., Wunsch, D.C., Domingos, H., "Device Degradation By High Amplitude Currents and Response Characteristics of Discrete Resistors." Work supported by DNA under Subtash R 990 AXE B097, Work Unit 42.

Tasca, D.M., (General Electric), Wunsch, D.C., (AFWL), "Computer Damage Characteristics Due to EMP Induced Transients," (to be published).

SECTION VI

DESIGN PRACTICES FOR EMP MITIGATION

6.1 INTRODUCTION

Protection of electrical and electronic systems from electromagnetic or induced electrical impulses is certainly not a new problem. The earliest efforts were probably associated with protection from lightning which is a natural phenomenon. As the propensity of electronic systems (radar and communications) increased, electromagnetic interference (EMI) became important and protection against its disruptive effects was required. In the 1960's the EMP from a nuclear burst was recognized as yet another potential disruptive source of electromagnetic energy.

There are many ways to protect systems against EMP. Many of the approaches and concepts were borrowed from EMC (electromagnetic compatibility) and lightning technology. Although this borrowed technology provides guidance for EMP mitigation; the EMC and lightning protective techniques, procedures, and devices are not adequate for EMP protection in most cases.

This section will deal with the protection philosophy, hardening techniques at the systems level, terminal protective devices, circumvention approaches, and quality assurance of the final product. As a reminder, the course deals with protection against the radiated EMP. Protection against the additional effects associated with the source region are not considered.

6.2 PROTECTION PHILOSOPHY

There are two kinds of degradation which must be protected against: (1) functional damage to critical portions or components in the system, and (2) operational upset of critical portions or circuits within the system. The type of hardening applied, and the level of protection provided, is a function of the sensitivity of the system to these disruptive effects.

- FUNCTIONAL DAMAGE
- OPERATIONAL UPSET

Protection is realized by choosing the most appropriate approach or combination of approaches, and the appropriate protection concepts, and then implementing these through design practices.

Protection Concepts

The subject of protection approach is concerned primarily with the level at which the required protection is to be achieved (i.e., the system level, circuit level, or component level). Applying protection at the system level, the primary objective is to keep the undesired energy out of the system, or at least the sensitive mission critical portions of the system. At the circuit level, the primary objectives are to limit the undesired energy reaching the sensitive circuit or components, and design less sensitive circuits. At the component level, the primary objective is to select the least sensitive component in terms of the undesired response while still meeting the performance criteria.

These approaches translated into concepts for providing protection against functional damage are:

For damage protection,

REDUCE:

- EMP EXPOSURE
- COLLECTION & COUPLING EFFICIENCY
- FRACTION OF APPLIED ENERGY
- COMPONENT SUSCEPTIBILITY

Operational upset is a circuit or subsystem problem. Therefore, in addition to the concepts for damage protection, additional techniques are available to minimize this effect. Stated as concepts, these are:

CONCEPTS FOR UPSET PROTECTION

- ALL DAMAGE PROTECTION MEASURES
- HIGH LEVEL DIGITAL LOGIC
- CODING
- HARD MEMORIES
- EMP EVENT SENSING
- SOFTWARE CIRCUMVENTION

These concepts can be implemented in a variety of ways, termed "protective practices." There are several viewpoints toward a rational, balanced, and complete consideration of protective practices. This section categorizes these design practices according to the level at which they are generally applied. It should be noted that some of the topics discussed are of concern at more than one level, although they are only presented once.

The protective practices categorization follows:

CATEGORIES OF PROTECTIVE PRACTICES

SYSTEM LEVEL
- ZONING
- CLUSTERING
- CABLE LAYOUT
- CABLE CONFIGURATION
- SHIELDING
- GROUNDING

CIRCUIT LEVEL
- PROTECTIVE DEVICES
- CIRCUIT DESIGN
- CIRCUMVENTION

COMPONENT LEVEL

DISCUSSED IN SECTION V

System Implications

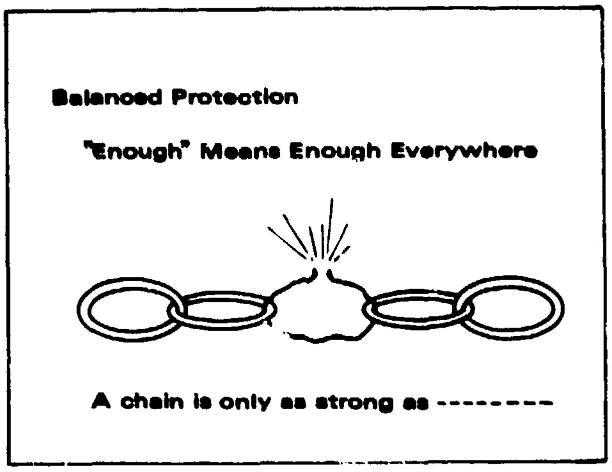
The decision to harden a system has both system cost and performance implications. These must be carefully studied before selection of a hardening concept or proceeding to the design phase.

In the past, systems have been categorized on the basis of size, type, user, or new versus retrofit, for example, and the hardening decisions based on these categorizations. Although these factors influence the protection design, the real point in making a hardening decision is the criticality of the system. That is, is the performance of the mission worth the price that has to be paid to insure success.

SYSTEM TYPES
- SOFT VS. HARD
- LARGE VS. SMALL
- MILITARY VS. CIVILIAN
- GROUND-BASED VS. IN-FLIGHT
- NEW VS. RETROFIT

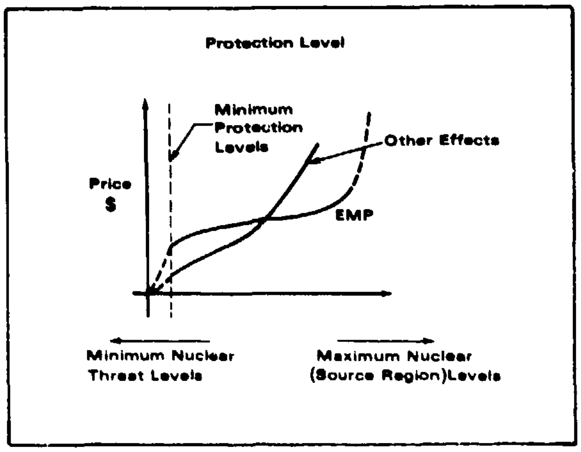
A BETTER CRITERION
- PRIMARY MISSION VS. PROTECTION
- SUCCESS VS. PENALTIES

Having made the decision to harden the system, the next consideration is the most cost effective approach. All systems are not vulnerable, and even those that are, all portions of the system are not equally vulnerable. Therefore, the same degree of protection is not required for all systems or portions of systems. In many cases, such as buried facilities, a complete envelope shield has been used to provide the basic protection and an electromagnetic boundary. This seems to contradict the previous concept of only hardening the mission critical subsystems. In those cases, however, the envelope shield was the most cost effective approach. In many cases, the more susceptible equipments will require additional protection which is provided by other techniques. Therefore, the various techniques for achieving the required protection must be studied and a balanced approach employing a combination of these techniques selected.



Perhaps the greatest single influence on EMP considerations is the cost of retrofitting existing systems and of properly designing new ones. Strangely, even to date, the cost patterns for EMP protection are not, as yet, well quantified; although some general trends can be indicated.

In the early days, it was not generally recognized that the protection-cost curve for EMP does not behave like those for other nuclear weapon effects. In all cases, the decision to protect represents a jump in systems cost, but beyond that, things are generally different, as seen here.



Some reasons for the cost curve behavior lies in the general properties of EMP protection hardware. The price factor in EMP protection lies in the introduction, or engineering of protection elements or hardware. Generally, making these simply "bigger" (or "thicker") does not increase the price as quickly.

For example, it costs little more to double the wall thickness of a shielded room. In fact, it may be cheaper because of the ease in welding thicker steel and lower union rates.

The cost of EMP hardening is highly dependent on the system and hardening requirements. Therefore, it is impossible to quote figures for hardening a system as a fixed cost. Further, many times the protection techniques utilized for EMP hardening also provide for EMC control and lightning protection. When this is the case, which it often is, there is no way to allocate the costs.

Cost information can be quoted, based on past experience, in terms of a percentage of the system design costs. For new systems where EMP hardening is considered from the onset of the design, the costs vary from about three (3) or five (5) percent to ten (10) percent of the system costs. For systems that must be retrofitted, the cost curve begins at approximately ten (10) percent and may go as high as 100 percent of original system cost. This indicates it is far more cost effective to EMP harden systems beginning with the initial system design.

Referring again to the cost curve, it is apparent that once the decision is made to protect a system, the degree of protection has only a small effect on the system cost. Because of the flatness of the cost curve, it is often desirable to design in a safety margin in terms of the amount of protection provided. This is desirable due to ambiguities in threat specifications, the variations in ambient environment, the uncertainties in manufacture and construction, and to counter the results of poor EMP field practice, maintenance and operational degradation. A relatively liberal margin is permissible; in the range of 10 to 20 dB.

To summarize, the incorporation of EMP protection into the design of a system certainly impacts cost and performance of the system. To do this in a cost effective manner, the system designer must: (1) identify the alternate approaches to achieving EMP protection for his system, (2) choose the optimum approach through good balance of protection practices, and (3) scale the hardness level to meet the protection requirements, including a reasonable safety margin, but which is well within cost restraints.

PROTECTION DESIGN

- IDENTIFY
- CHOOSE
- SCALE

Intercommunity Relationship

EMP protection practices are based on on the same basic concepts followed in the Electromagnetic Compatibility (EMC), Electromagnetic Interference (EMI), Hazards of Radiation to Ordnance (HERO) communities.

The state-of-the-art in EMP hardening is still evolving, whereas the state-of-the-art in the related communities has evolved to a point where complete and thorough quantified design practices, specifications, standards, and quality control procedures now exist. Much of this technology and documentation, although useful as a guide, is not directly applicable to the EMP problem.

Quantified Existing
Design Practices and Quality Control
Not Always Appropriate
for EMP Problem

The related communities are concerned, in many instances, with mitigation of undesired effects at the subsystem or equipment level. The nature of the EMP and its interaction with systems is such that mitigation must initiate at the system level. Consequently, protection against EMP effects is very often system specific.

6.3 HARDENING DESIGN PRACTICES

The discussion of design practices for EMP hardening is subdivided into two major areas: (1) the Systems Aspects, and (2) the Subsystem/Circuit/Component Considerations. The Systems Aspects deal with those aspects which must be decided at the systems level and must be uniform throughout the system. The Subsystem/Circuit/Component Considerations deal with the intrasystem aspects and pertain to protection at the subsystem and equipment level.

Systems Aspects

In considering system hardening against EMP, it is necessary to initiate hardening design at the systems level. It is at this level that the system configuration, intersystem communications and data transmission configuration, shielding philosophy, and grounding philosophy are determined.

It is important to keep in mind that, when we say "system," this can encompass a broad spectrum of configurations in size, shape, and complexity. A pocket transceiver is as much a "system" as is an ABM radar site. Thus, its definition is: a complete, self-contained primary-mission entity.

To establish a common language for discussion, a few definitions are in order:

SYSTEM LEVEL CONCEPTS/DEFINITIONS

- SYSTEM
- ZONING
- CLUSTERING
- LAYERING
- DAMPING
- VIOLATIONS
- FIXES

Some other system-level concepts and definitions are noted here:

Zoning:	The identification and integration of regions of similar EM environment and/or susceptibility.
Clustering:	The grouping of elements of similar characteristics and purposes.
Layering:	The sequencing of zones and protective measures between outer environment and inner equipment.
Damping:	The use of lossy elements or materials to absorb EM energy.
Violations:	The features which represent defects from a systems hardness viewpoint.
Fixes:	The measures taken to rectify violations.

System Geometry and Configuration

The first step in hardening design is to identify and allocate the hardening for various subsystems and equipments which comprise the system. Zoning is one approach for achieving this goal.

Zoning

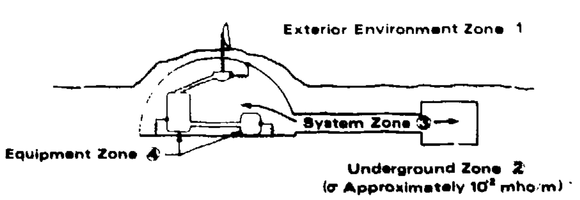
Electromagnetic zoning may be established in two ways:

(1) Environmental zoning, in which the magnitude and shape of the field pulse are defined within the successive regions from the outside in.

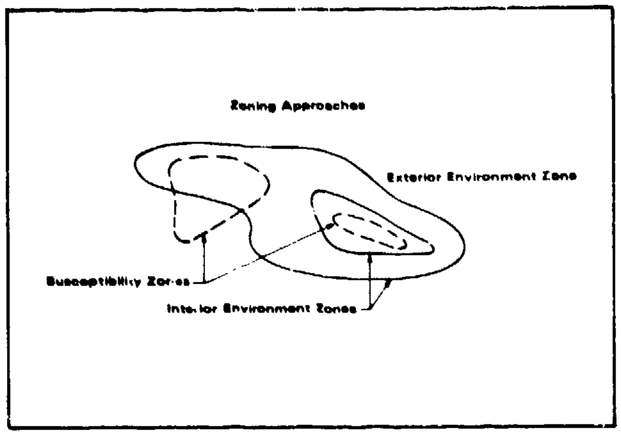
(2) Susceptibility zoning, in which the magnitudes and frequency (or time) domains corresponding to the vulnerability thresholds are scaled from the inside out.

The choice of which approach is followed is often dictated by other factors. For example, if a system already exists, it is often easier to establish the zones by the first approach (environmental zoning). The reason for this is that the system configuration, to a great extent, is fixed. This is depicted for a buried system where zones 1, 2, and 3 are established by the construction technique. An additional zone, 4, may be determined by either approach. Either approach is also applicable to new designs where equipments may be grouped by their susceptibility levels, thus establishing the system configuration and necessary EM zones.

System Aspects



A well designed system will exhibit an appropriate coincidence between environmental and susceptibility zones. An example of "bad" zoning is indicated where the susceptibility zone crosses an environmental zone boundary.



It is common practice to define the levels of different zones in terms of relative dB. Zone boundaries are usually associated with physical features such as walls, bulkheads, compartments, cabinets, etc.

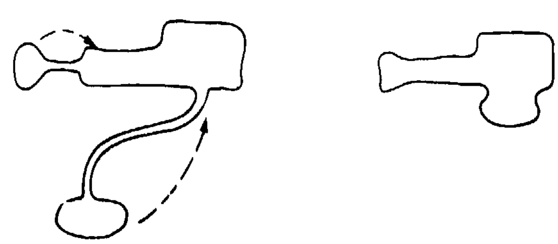
Zones may also be delineated in terms of other electromagnetic interaction specifications. Examples of some of the more demanding requirements are the EMC specifications, Mil-Std 461/462/463, and the TEMPEST standards, NACSEM 5100 series.

Clustering

Evidently one of the things which should improve EMP hardness is the reduction of the area over which vulnerable elements are located. All other systems aspects being equal, it is generally best to contour the EMP zones as compactly as possible.

This is especially important if upset, such as computer memory erasure, is concerned.

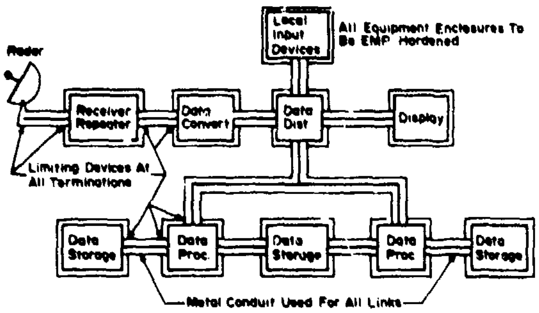
Clustering



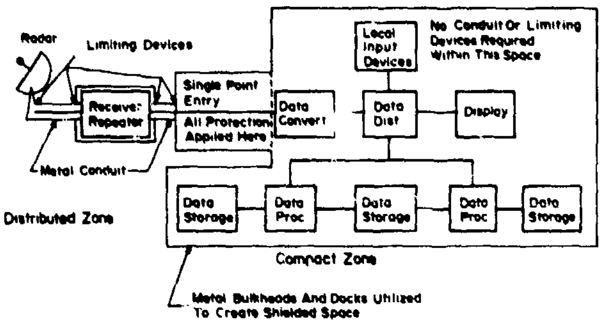
"Gerrymander" Zone

Cluster Zone

While the idea of a small single zone is desirable, it is not always feasible. For systems that have distributed elements at a variety of locations, it is often necessary to implement a gerrymander type zone maintaining zonal integrity (shielding) via well shielded cables, such as conduit. An alternative is to cluster various portions of the system and minimize the cabling, or isolate the various zones.



a) Totally Distributed System Protection Scheme



b) Partitioned System Protection Scheme

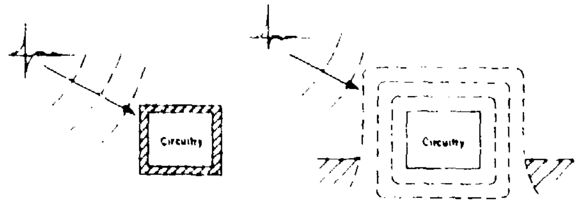
EXAMPLE OF SYSTEM PARTITIONING

Layering

Most of our simple examples here show EMP protection as appearing in several successive geometric stages or layers. Of course, each boundary has to be complete, in the sense that apertures and penetrations must be treated to preserve what was gained at that layer.

There seems to be a tendency to deal with EMP at one or two boundaries. In many EMP cases, this is quite unrealistic. For instance, in a deeply buried system, it is plainly obvious that some protection can be gained almost "free" from the earth cover itself. It is also unrealistic in "porous" systems -- that is, systems with very many apertures and penetrations.

Layering



Single "Thick Fence" vs. Multi-layer

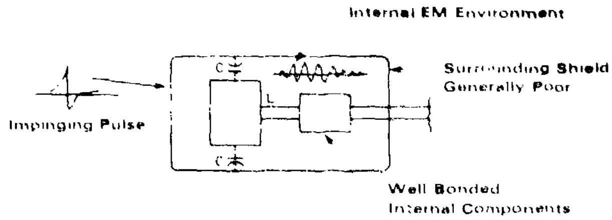
Ringing

There are two approaches to EM field protection. One of these is the "iron curtain" method, in which the various elements are thoroughly shielded and electrically isolated from one another. The other is the "common sink," in which the various elements are massively connected together.

The difficulty is that one cannot do either thing thoroughly. Elements must be connected together somehow, but they cannot all be placed in intimate contact. The result is something in between, which often acts like a high-Q EM cavity or LC circuit.

This is basically why many partially shielded systems exhibit strong ringing when excited by means of an EMP simulator.

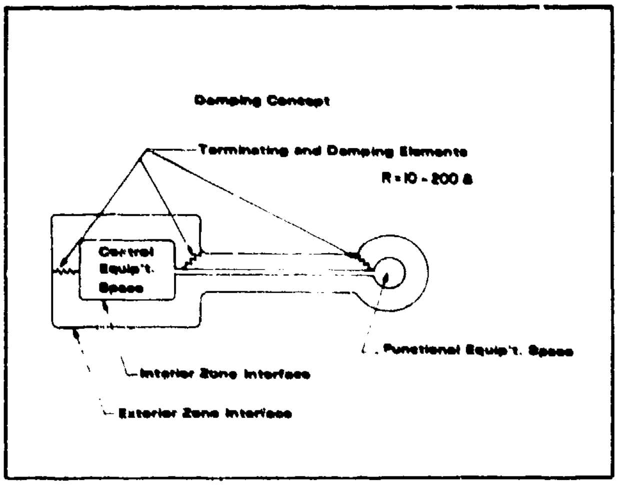
Ringing



Such ringing represents efficient storage of EM energy and a prolongation of the time during which it can be coupled to internal elements and circuits. This energy storage can be reduced by spoiling the Q of the enclosures and circuits. The concept is illustrated here by the insertion of parallel damping resistors. (Series damping requires careful circuit analysis to avoid making matters worse).

This technique has been very successful in some types of nuclear test interference problems. Most shielding systems have characteristic impedances in the range of 5 to 200 ohms. Damping resistors of corresponding value are used; the exact value is not critical.

This technique is most appropriate where the shielded enclosure is (for a variety of reasons) poor, permitting entry of the higher frequency (ringing-frequency) components.



Violations and Fixes

The zoning concept has another advantage in complicated system evaluations. It permits the definition of specific locations and components (along a boundary) requiring EMP treatment. In the strictest analytic sense, one assigns a minimum dB margin which all points and elements within and at such a boundary are to satisfy. Those that do not are at once identified as "violations." As mentioned previously, "fixes" clearly encompass those measures taken to redress these situations, or, in some cases, to redress their consequences.

Violations generally fall in one of four broad classes as outlined here.

Violations and Fixes

- **Distributed couplings (zonal field)**
- **General transparency (finite sheets)**
- **Apertures (singular holes, seams, etc.)**
- **Penetrations (insulated conductors)**

In assessing a system, it is easy to overlook conductors which are not labeled as electrical circuits -- or even labeled at all. For example, long lengths of rebar (and sometimes tubing) used in structural piling. Here are listed some other conductors which have turned out to have possible EMP significance, in terms of providing EMP collection and/or coupling paths into otherwise protected systems.

Past Surprises

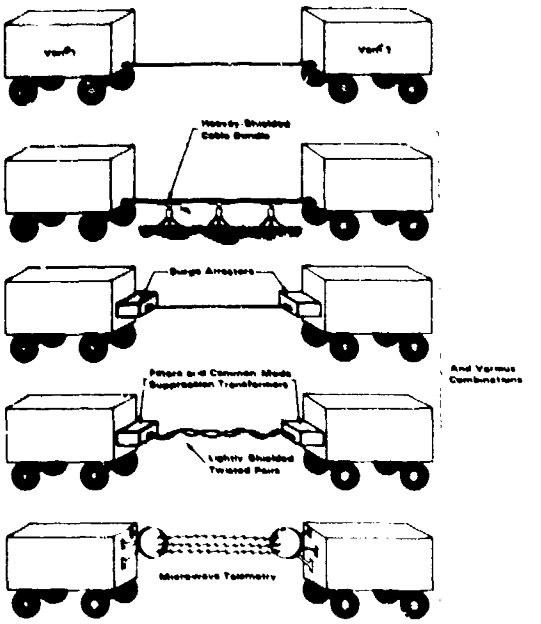
- **Service Features**
- **Miscellaneous Electrical Features**
- **Construction Features**

Communications and Data Transmission

In any system, it is necessary to provide communication and data transmission between various portions of the system. This must be accomplished without violating the zoning configuration established for the system.

The development of EMP criteria for these links does not necessarily start with the de facto system connection diagrams. Rather, it should start in the determination of what is to be connected to what, and how this is to be accomplished. As we will shortly indicate, much may be done to ease the hardening problem by more judicious circuit management. Of course, once the inescapable connection requirements are determined, then one must get to the specific hardware issues.

There are several system configuration options available for providing the necessary communication and data links. Any of the options depicted can provide the required EMP hardness.



Hardware design, such as cable shielding, terminal protection devices, etc., required to implement these options is discussed later in this chapter. At the system level, it is mandatory that the hardening approach be specified since these options are very difficult to combine. Basically, the options indicated can be grouped into two major categories: (1) the preservation of zonal integrity through extension of the zonal boundary via extremely well-shielded cabling, or (2) preservation of zonal integrity by isolating the zones via terminal protective devices, or by non-conducting transmission links.

Shielding

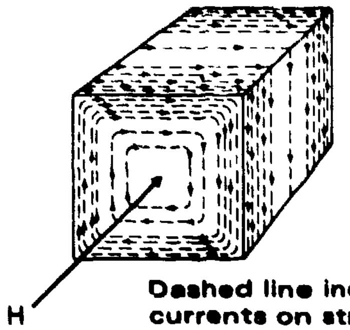
As indicated previously, there is considerable identity between zoning and shielding. Indeed, when "good practice" alone sufficed, it centered on shielding as the controlling feature. Questions of circuit layout, cabling, and filtering tended to be subordinated. Here, shielding is treated as simply one aspect of a larger formation.

For comparative purposes, it has become customary to rate a design or product in terms of "shielding effectiveness," the attenuation of the fields stated as a function of frequency. The shielding effectiveness will depend on the size of the box, the location of the item, the frequency domain and the method of measurement. Basically, it can be viewed as a measure of a certain internal environment "with" vs. "without" the protective scheme.

In the EMP frequency domain, a dominant mechanism in shielding effectiveness is inside cancellation or field reflection due to induced surface currents as illustrated here. Thin walls, high resistance paths, apertures, seams, etc., seriously affect the reflection or cancellation characteristics and serve as internal field generators as well.

A good shield must, therefore, be sufficiently thick, continuous, complete and tight. This is essential for shielding above 60 dB $(S.E.(dB) = -20 \log \frac{Hi}{Ho})$.

How Do EMP Shields Work?



Dashed line indicates induced currents on structure

Current distribution on a box-like enclosure caused by a low-frequency magnetic field

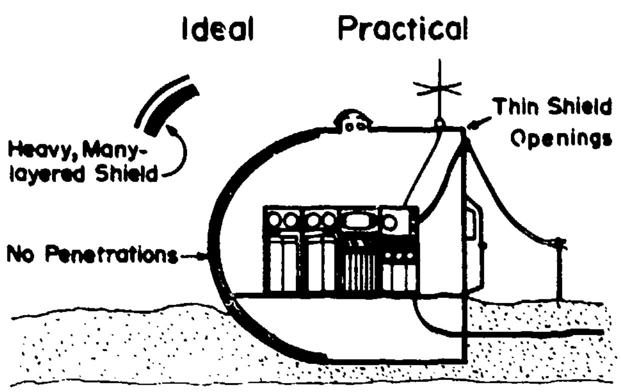
Of course, mechanical and electrical inputs and outputs are also essential. Economic realities place real limits on wall thickness.

So, shielding hardware considerations generally boil down to compromises in relation to:

Wall thickness and material

Apertures -- tightness

Penetrations -- conductors.



Wall Thickness and Material

In the EMP time domain, the dominant mechanism in shielding is the induced surface current. This is concentrated in a surface layer, the "skin depth (δ)," is given by

$$\delta = \frac{\tau}{\mu_o \mu_r \sigma}$$

where

τ = time

μ_o = $4\pi \times 10^{-7}$ permeability of air

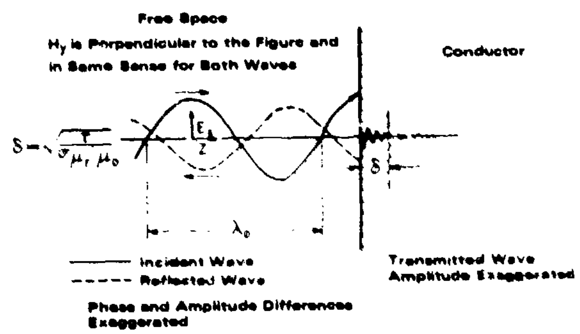
μ_r = relative permeability of the material

δ = conductivity of the material

Since the skin depth varies as $\sqrt{\tau}$, it is difficult to significantly reduce the internal $\dot{B}$.

In most practical cases, it becomes difficult to justify a shielding thickness much greater than that required by mechanical strength and rigidity.

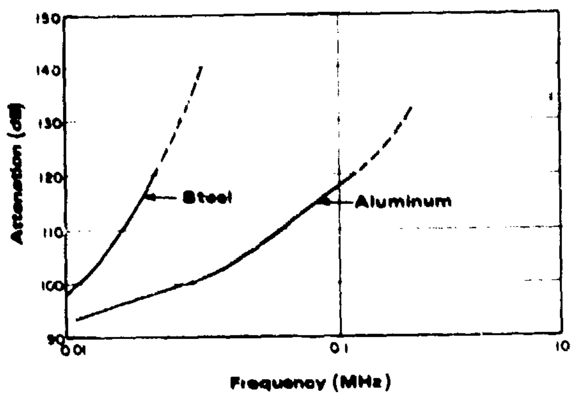
Wall Thickness



Schematic Representation of the Reduction in Amplitude of an Electromagnetic Wave on Entering a Metal

The skin depth also depends on $\sqrt{\mu\sigma}$ ($\mu = \mu_o\mu_r$). Hence, there <u>is not</u> as much difference between copper and steel as one might think. We see here that the main advantage of steel is to obtain the same attenuation at about one order of magnitude lower frequency. Note also the large attenuations realized for relatively thin sheets. These values are for infinite sheets of material.

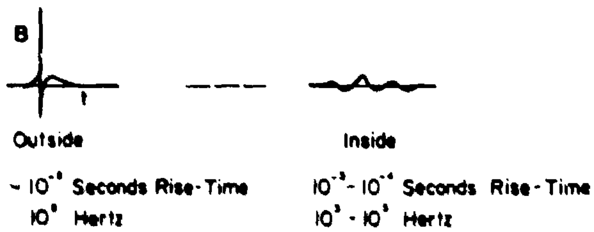
Materials



Magnetic Attenuation versus Frequency.

As indicated, the attenuation (shielding effectiveness) for magnetic fields is a function of frequency. The higher the frequency (the greater the $\dot{B}$), the greater the attenuation. Therefore, the shield tends to be a $\dot{B}$ reducer.

A SHIELD = $\dot{B}$ REDUCER

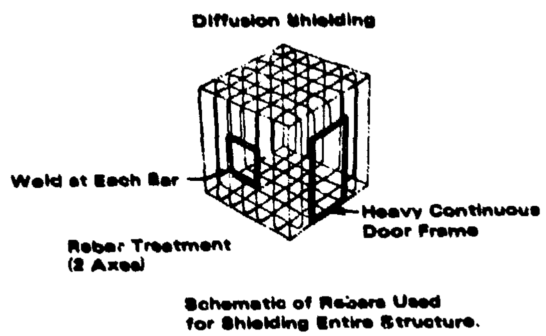


A further implication is the lower the frequency the less the magnetic field attenuation. Therefore, good low frequency shielding requires the use of very high permeability (μ) materials, such as hypernom and conetic, or very thick materials such that magnetic field ducting is realized.

Diffusion Shielding

There is another kind of shield -- the semipermeable type; examples of which are earth cover and rebar grids. Here the effective skin depths may be large (δ = 28m for earth with conductivity of 10^{-2} mhos/m at a frequency of 100 kHz), and the total attenuation relatively small -- about 30 dB. Usually this is used in combination with smaller, internal, and more complete shields. Often such a shield appears as a zone enclosure of opportunity, such as in buried or heavily reinforced structures.

The waveform appearing inside such a diffusion zone will generally be a combination of a short spike (possibly associated with apertures) and a longer "tail," related to the induced skin currents on the conductor.



Schematic of Rebars Used for Shielding Entire Structure.

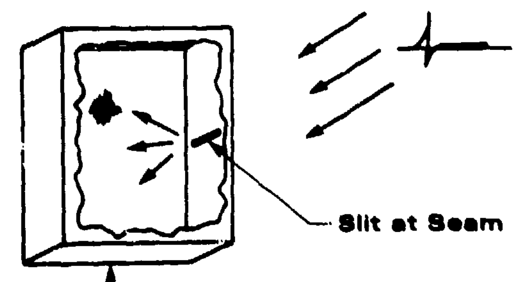
Apertures

There are many different kinds of "apertures." They may be divided as intentional and as unintentional. The single worst class of violations of good EMP protection practice is found in the accidental or unintentional compromise of shielding integrity. Anything which interrupts the skin current path on a shield increases its impedance and acts as a radiator into the internal region. Hence, the effect of a seam crack is not measurable simply by its physical area, which may be quite small. If it is near a region of high surface current concentration, it can couple energy to the interior many times greater than you would superficially guess. In particular, physical breaks -- such as seams and bonds, however well made -- represent a constant threat to integrity and protection value.

Of course, it is almost impossible to fabricate a shield as a single, unbroken, electromagnetic enclosure. Large system enclosures can only be constructed by assembling large numbers of sheets or plates. Technically, the contact lines or seam between such single pieces represent potential apertures.

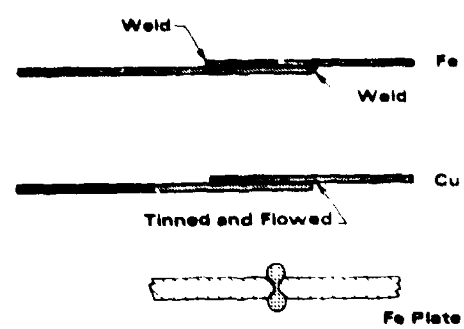
Depicted is an "idealized aperture"-- a long slit in a shield wall due to a poor panel joint. It behaves approximately like a slot dipole antenna. In effect, it is excited by the EMP fields and the induced skin currents and radiates a wave, characteristic of its length, into the box.

Apertures



The most common large-scale seam fabrication techniques involve welding for steel or aluminum and soldering or brazing for copper. These fabrication methods in themselves place certain minimum thickness criteria on the material. Thin sheets tend to "burn through" during welding. Therefore, at least two mechanical aspects impose minimum thickness requirements, which may be greater than required by the EMP criterion: strength and fabricability. Such thicknesses run from 60 to 300 mils for medium-large construction. Overlap should preferably be 10-20 times the sheet thickness for thin sheets. Butt joints can be acceptably used for thick plate, but this usually requires welds on both sides, with careful probe tests for weaknesses.

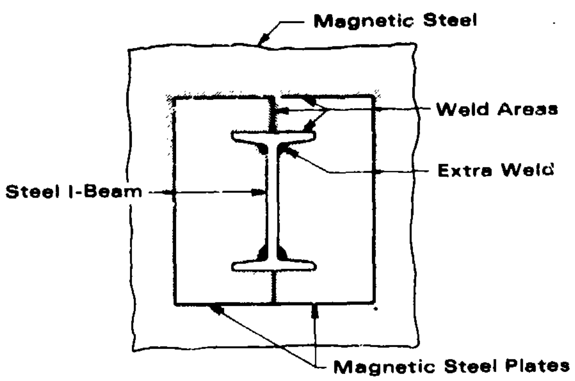
Seams



The necessary mechanical thickness provides an implicit (and high) protection level. Inexpensive assembly methods, such as tack welding, may seriously erode the protection level due to aperture leakage at the open seams throughout the structure. The "good shielding" criterion may require continuous and meticulous welding along all seams in order to match the protection value inherent in the material itself.

As an example of the welding integrity criteria, an ICBM test facility had a shielded room built into its base. This enclosure had a 100 dB requirement in the UHF domain. The construction was such that several large structural I-beams passed through the room. In order to "seal" the room electromagnetically, specially cut plates were welded around the beams and onto the steel walls. These seam areas were tested by using a transmitter loop outside and smaller receiver loop probe inside. A single continuous welding pass proved inadequate to prevent leakage at this seam. Several additional passes were needed around these beams, both inside and out. These were made in such a way as to build up a thick weld. The inner corners were particularly difficult points, as indicated here.

Weld Integrity



Gaskets and Bonds

Considering the difficulties encountered with such seemingly "tight" apertures as welded seams, it is no surprise that metal-to-metal contact surfaces, held together by simple mechanical pressure, can constitute serious violations of shielding integrity. Such contact areas are unavoidable at functional apertures, e.g., access doors, service hatches, equipment panels, etc.

There is extensive literature on all manners of bonding long, continuous, metallic, contact lines. They deal with a range of bonding permanency, from permanent, once-made joints, through rarely-disturbed service panels, to continuously exercised doorways. Of course, the latter represent the most difficult problem in dependability and maintainability.

The basic mechanical requirements for simple reliable seam bonds are absolute flatness and electrical cleanliness. Neither of these is generally achievable in other than ideal laboratory conditions. The pragmatic hardware problem is then to obtain low-impedance continuous contacts at an acceptable level of "dirtyness" and "deformation."

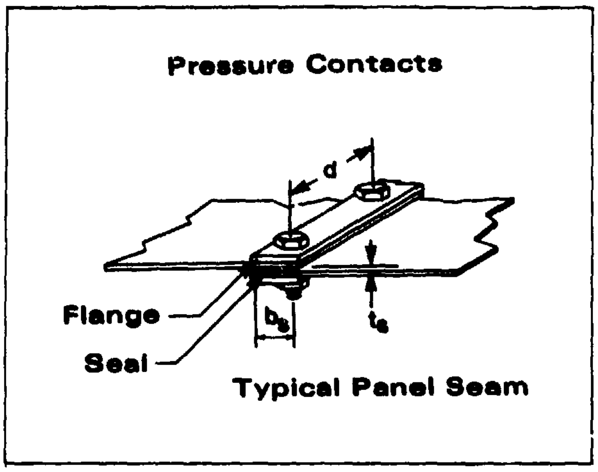
Electrically clean surfaces can be readily obtained with pure tin, gold, palladium, platinum and silver! Zinc, cadmium, and very thin gold platings are considered as acceptable substitutes. Easily oxidized materials (like aluminum) should be avoided. Lubricants are capricious. In some cases, they will inhibit corrosion and oxidation and facilitate good metal-metal contact. However, motor oils, for example, are more apt to do just the opposite.

Controlled roughness (machine sining and knurling) is generally better than attempting to achieve a smooth surface for mating parts. When controlled roughness is utilized, the total contact area is generally greater, and easily predictable, than for smooth surfaces which mate usually at only three points (no surface is perfectly smooth).

CLEAN CONTACTS

GOOD		BAD
TIN, ZINC, CADMIUM PLATINGS (ALSO GOLD, PLATINUM, AND SILVER)		ALUMINUM, IRON (OXIDES)
THIN ←	PLATINGS	→ THICK
SPECIAL ELECTRICAL ←	LUBRICANTS AND PAINTS	→ SULFUR-BASED AND OXIDIZED
SCORED, KNURLED ←	SURFACE	→ MILL-SMOOTH

Roughness is one way to compensate for surface irregularity. An ultimate way to do this is to use deformable conductive gaskets. A good way to understand the pressure contact problem is to consider a panel seam, bonded by means of bolts and flange strips, as illustrated here. In the frequency domain of interest, seams of this type require specific contact pressures of 60 to 100 pounds per lineal inch for 80-100 db attenuation. Obviously, this form of seam is best for "once-only" cases, which are expected to be broken very rarely, if ever, during the system's life.



People have also resorted to the "gasket" solution for "bonding" peripheral contacts which would only be occasionally broken. It is also useful for irregular or deformable surfaces.

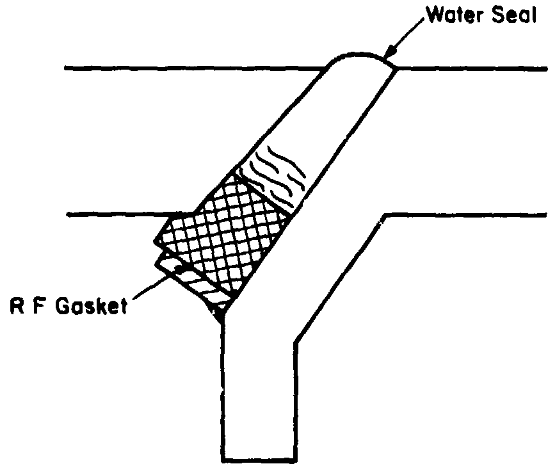
There are two "fairly" good types:

1. The flat molded metal gasket which deforms slightly under pressure. This is a "throw-away" in the sense that it cannot be reused.

2. The braided cord gasket. A variety of exotic designs appear on the market. The good ones from an attenuation standpoint utilize deformable metal cores. Unfortunately, these have low resiliency and can only be reused two or three times. The synthetic core, double braid gaskets are generally more transparent in a given geometry. The single braid types provide even less attenuation. Braided gaskets are not generally recommended for exposed, unmaintained situations for EMP protection.

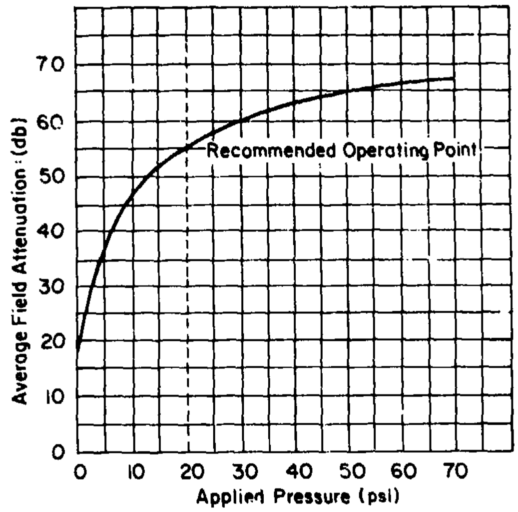
Corrosion control of the gaskets and the associated mating surfaces is also a problem. Considerable maintenance is required to insure good electrical contact.

Depicted is a typical application of a knitted mesh gasket for a ship missile loading hatch. Note that the RF gasket is protected from corrosion by a water seal on the hatch.



MISSLE LOADING HATCH RF SHIELDING (CLG), (METHOD A)

The average field attenuation that can be obtained with knitted wire mesh gaskets as a function of the applied pressure is indicated here. It should be noted that application of higher pressure increases attenuation, but shortens the lifetime of the gasket as it permanently deforms.

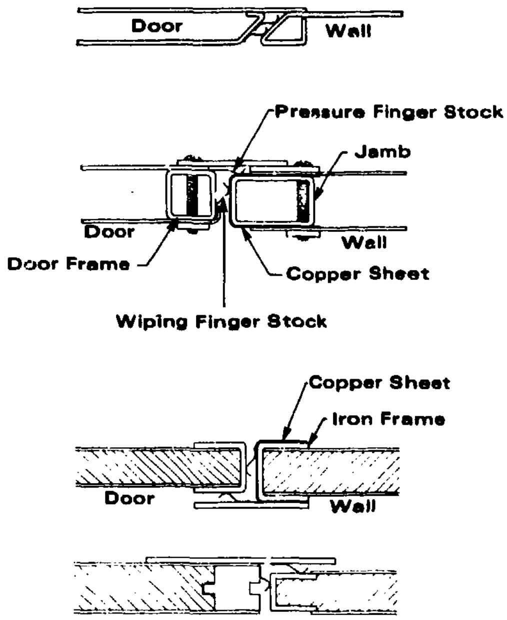


KNITTED WIRE MESH GASKETS, AVERAGE FIELD ATTENUATION

Resilient finger stock is a favorite solution for doors and hatches which must be frequently used. Finger stock should be used in double rows. Some people suggest that the rows should be staggered for maximum attenuation so that the fingers in one are opposite the slots in the other. At the higher frequencies, this seems reasonable when one considers the radiation pattern of each slot, seen as a tiny dipole.

Finger stock is probably the most difficult protection hardware to maintain. Traffic inevitably brings with it dirt and abrasion. The doors and frames must be extra stiff if the fingers and the contact surfaces are to maintain their register.

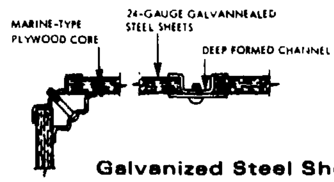
Finger Stock and Doors



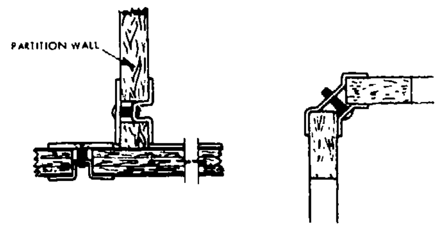
The joint constructions illustrated here show some of the ingenious ways in which shielding engineers have solved the problem of structural flexibility with reliable shielding effectiveness. They typify shielded enclosures for R.F. testing and measurement. At present, at least, it is not likely that military systems would employ such components except as accessories in production and testing phases.

The prefabricated bolt-together enclosure has enjoyed wide acceptance. It does, however, require periodic maintenance. The frame shifts cause open slits and metal-to-metal seam corrosion. Where high shielding requirements exist, serious consideration should be given to the welded seam enclosure.

Shielded Enclosures



Galvanized Steel Sheet with Plywood Core.



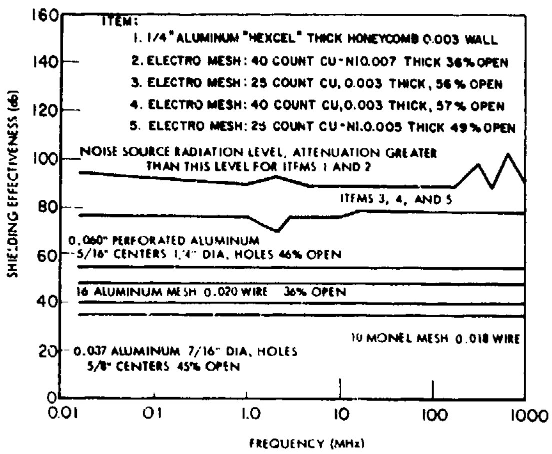
Partition Joint Example.

Open Apertures

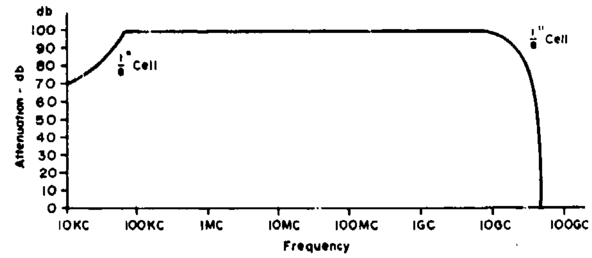
So far we have dealt with apertures which could be "closed" electromagnetically by means of conductive materials (sheets) and construction similar to the surrounding shield. The significant problem was with peripheral control. Some mechanical requirements, however, call for a physically open aperture for such things as ventilation, microwave lines, etc. Two broad classes of "solutions" are common for these: screens of various types, and "waveguides-beyond-cutoff." The latter can also be used sometimes for entrance passages and doorways to avoid the finger-stock problem, where penetration of high frequency content is clearly not a problem.

Ordinary heavy-duty screening can provide on the order of 40 dB attenuation. The trouble with ordinary screening lies in corrosion and oxidation which can break the contact between individual wires. Electromagnetically, an old piece of screening may be a good coupler. The specially fabricated materials like "electromesh" are treated to resist this action.

Screens

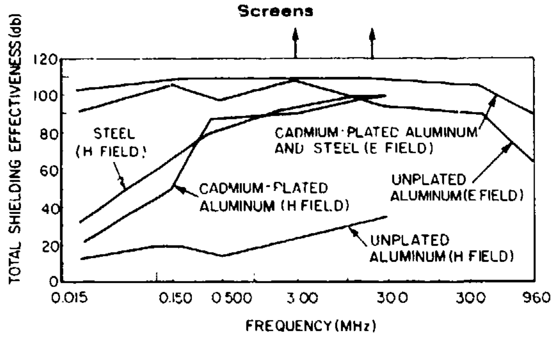


attenuation of various screening materials



PREDICTED THEORETICAL ATTENUATION

Screens

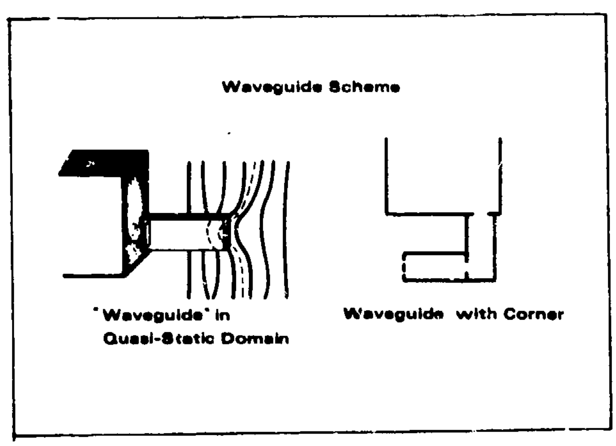


claimed performance of an average commercial honeycomb filter

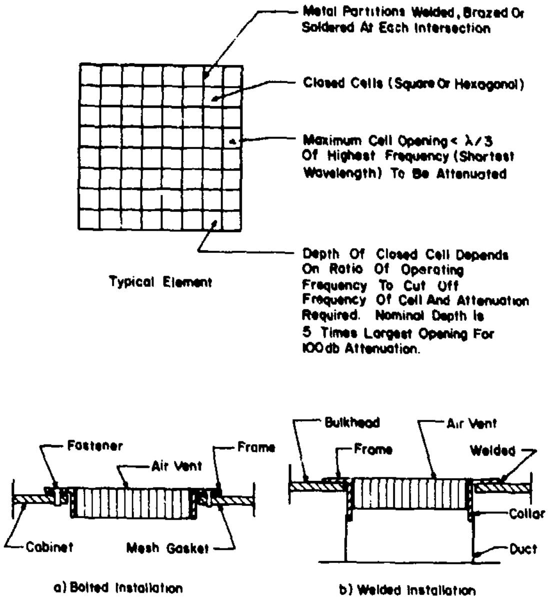
Hexcel is usually satisfactory, but there have been instances of poor quality control in which the glue between the foils acted as an insulator. True "honeycomb" screening provides the best compromise between shielding and air flow. Where air flow is important, best results are obtained with honeycomb which has soldered, brazed, or welded contacts between foils.

The "waveguide-beyond-cutoff" is somewhat of a misnomer. Over most of the EMP frequency domain, such a geometry is really behaving more like a quasi-static "field-bender." The idea is to design it so that its cut-off frequency is significantly well above the high-frequency "roll-off" in the environmental spectrum. This is not difficult to do if it is under many feet of earth or it is already protected by some partial attenuation, such as a welded rebar cage. These situations tend to move the roll-off to lower frequencies, as we have indicated before. For example, a two (2) meter high driveway would have a cutoff frequency of 75 MHz ($f_c = \frac{c}{\lambda}$, where $\lambda = 2a$ and a is the maximum doorway dimension). For frequencies of less than 30 MHz, the attenuation is between 12 and 13.65 dB per meter length within the guide.

The approach is fine for ventilation, but care must be taken not to make the guide a propagating structure by running cables through it.

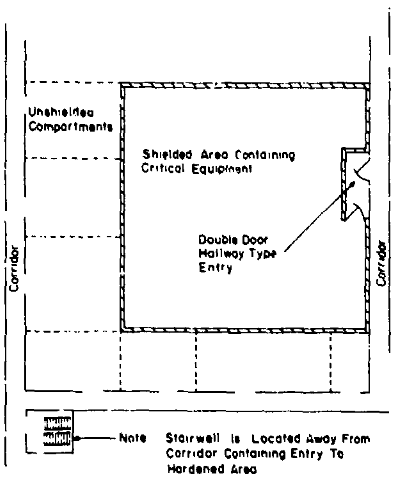


The use of the "waveguide-beyond-cutoff" approach for ventilation shafts in structures is one typical application. High frequency roll-off is obtained by partitioning the overall opening into an array of smaller openings as depicted. This will cause some reduction in airflow depending on the effective area reduction of the overall opening.



USE OF HONEYCOMB MATERIAL FOR SHIELDING AIR VENTS

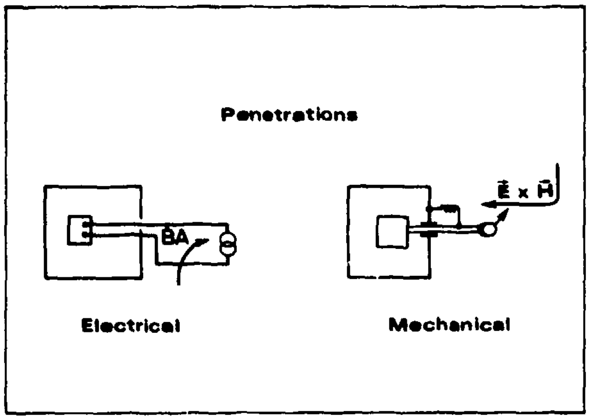
Another technique to handle large apertures, such as personnel entry ways into shielded compartments, is through the use of a protected entry. Shown is a technique using a double shielded door hallway type configuration. Interlocks should be provided if critical equipments are housed in the compartment so both doors cannot be open at the same time.



DOUBLE DOOR HALLWAY TYPE ENTRY TO HARDENED AREA

Penetrations

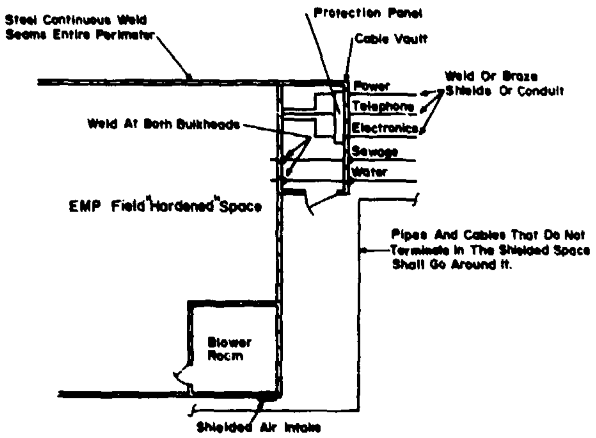
There are many kinds of "penetrations." Most commonly, one thinks of an insulated conductor passing into a facility or system. It may be carrying power or functional signals, but uninsulated, "grounded" conductors, such as motor shafts, can also represent penetrations. Thus, there are two broad classes -- electrical and mechanical. We see here ways in which each can provide paths for coupling and transferring energy from the external to the internal zones. Note particularly that mechanical penetrations can be deceptively protected by innocent-looking bonds, which are really high-impedance couplers.



The existence of a true "electrical penetration" corresponds to an intentional or unintentional violation of the zoning concept. If an electrical circuit is carefully confined to a single EM zone, then its penetration through a shield does not, in fact, constitute a violation. In principle, it cannot transfer energy which would not be there in its absence. We make this seemingly simple point to emphasize the necessity for observing zonal hierarchies in providing conductor and cable shielding (as discussed in a succeeding section).

But what about unavoidable conductor penetrations -- such as, for instance, long wire antennas? One thing to do to rectify such situations is to provide entrance protection in the form of filters or active devices (zener diodes, spark gaps, etc.). These are discussed in the section on "Protective Devices." These protective devices should be located in vaults or small shielded boxes.

Since all external conductors are collectors of EMP energy, they must all be terminated at the entry to the shielded compartment (or equipment cabinet). If these terminations are allowed on a haphazard basis, (terminating on all sides of the enclosure), they will inject additional current on the enclosure resulting in higher fields on the inside. Therefore, to keep this energy from entering, all penetrants are brought into the protected space via a single point of entry into an entry vault. The conductors (pipes, condui'?, cable shields, etc.) are peripherally welded to the enclosure at the entry point. All protective devices are contained in the vault and the output electrical leads into the enclosure protected by feed-through capacitors, for example. These terminal protection approaches are discussed in subsequent paragraphs of this section.



TYPICAL EMP HARDENED SPACE

Finally, one can isolate that portion of the system (which really goes back to systems and circuit layouts) and simply make its terminal circuits very hard.

So we see that the treatment of purposeful electrical penetrations is not really a "shielding" topic.

ELECTRICAL PENETRATIONS

NOT REALLY A "SHIELDING" HARDWARE PROBLEM -- WHY?

ZONE VIOLATION ⟶ GO TO "SHIELDED"

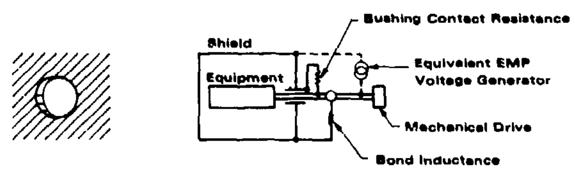
ENTRANCE PROTECTION ⟶ GO TO "PROTECTIVE DEVICES"

TERMINAL HARDENING ⟶ GO TO "SYSTEM ASPECT" AND "COMPONENT SUSCEPTIBILITY"

We noted before that a conductive metallic penetration may be deceptively protected. Consider, for instance, a shaft passing through a bushing. In the case of a metallic shaft with a conductive bushing and a bond strap, it can be treated analytically as a parallel R L circuit as shown. Simple circuit analysis can be applied to determine the fraction of the current which is not shunted by the bond strap or the bushing resistance. If the bushing has a high contact resistance (or is an insulating bushing), at high frequencies ($WL \geq R$) the shaft can act as a probe antenna coupling directly to the interior as a skin current or by reradiation. This type of penetration could easily result in a 30 dB lead in a 60 dB shield.

In the case of a nonconductive shaft, the configuration can be modeled for analysis as a dielectrically loaded circular aperture. The aperture problem becomes important when the aperture diameter is equal to or greater than the half wavelength of the highest frequencies associated with the incident waveform. For small shaft dimensions (5 to 10 cm), the coupling through the aperture would be small (a few dB) over the frequency spectrum of the EMP.

Mechanical Penetrations



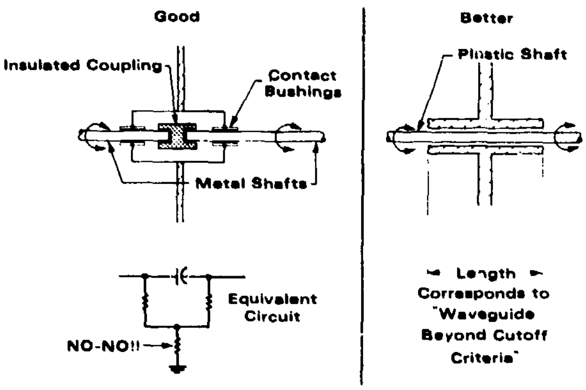
The principle here is to "rectify the obvious." We will discuss some embodiments for these first two types of fixes.

Treatment

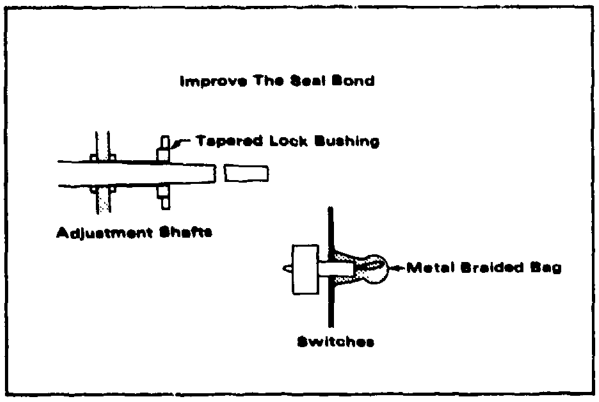
- **Break the metallic path**
- **Improve the seal bond**
- **Examine electromechanical servos, relays (plus filters on the wires)**
- **Examine non-conductive, exotic schemes**

The first approach is to break the metallic path. This treatment replaces the continuous conductor by a small capacitance. Its equivalent circuit represents it as a filter between the two EM zones. Note that the separate small enclosure must be well-bonded to the shielding partition between the two zones. An impedance at this point turns it into an effective coupler. This is equally true for filters and for other enclosed protective devices located at a partition between zones. An alternative approach is the use of the waveguide-beyond-cutoff technique. This approach requires the mechanical shaft to be non-conductive. In sizing the waveguide, it must be analyzed as a dielectrically loaded guide with the dielectric being the mechanical shaft.

Break the Metallic Path



Shown are two examples of unusual treatment. A simple way to improve a leaky manual adjustment shaft is to mount it through a split-bushing with a taper-thread locking nut. When the locking nut is tightened, a low resistance bond is achieved between the shaft and the equipment panel.



In "explosive proof" hardware, or in equipments with very high shielding requirements, a simple fix is to utilize a braided metal bag to increase the shielding as shown.

Grounding

The primary purpose of grounding is the protection of personnel and equipment. Electrical codes require that electrical and electronic equipment cabinets/frames be connected to the surrounding media (building, earth, etc.) in such a manner that no shock hazard exists due to a voltage difference between the equipment and the surrounding media. For equipment protection, the purpose of the ground is to provide a fault current path so sufficient fault current can flow to cause circuit breaker or fuse actuation. Ground further prevents the buildup of electrostatic or transient voltages that may cause insulation damage. Lightning protection is a special case of transient voltage buildup which is shunted to ground, in most cases, via surge protective devices.

In other words, the purposes of grounding are to provide an equipotential connection between equipments and surrounding structures/media, and a return current path for fault currents. For signal circuits, two-conductor transmission lines are used. Tying the signal circuit to ground prevents electrostatic drift and provides a common reference provided the ground circuit is truly an equipotential plane.

The discussion which follows will divide the grounding problem into earth grounds ("exterior" grounds), and equipotential or reference nodes ("interior" grounds). Emphasis will be placed on meeting system grounding requirements without becoming a major source of EMP pickup.

Both Interior and Exterior Grounds are Needed for Other System Reauirements

Both Interior and Exterior Grounds can be a Major Source of EMP PICKUP

Earth/Exterior Grounds

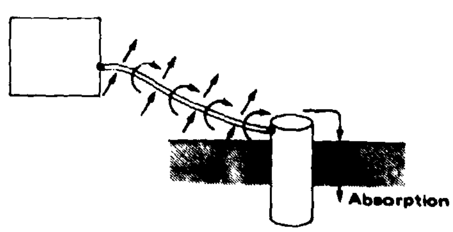
There are at least three reasons for considering how EMP and exterior grounds interact. First, there are long wavelength threat components with which ground circuits can meaningfully couple. Second, system grounds are essential for any number of other reasons; hence, their EMP coupling is a germane issue. Third, it has been pragmatically established that grounds make a noticeable difference,-- good and bad -- in nuclear test and nuclear simulation instrumentation.

Of course, a grounding system can be put to advantage in EMP control. But this needs to be integrated with other grounding requirements, i.e., lightning, power, etc.

The basic idea of earth grounding is to provide an equipotential distribution between the structural members of a system and the surrounding natural environment. This concept is perfectly valid only for the ideal case of static fields, infinite earth (ground) conductivity, and no current flow. Thus, earth grounding is an attempt to connect, in a field-significant way, to the large, but poor, conductor. In the cases of shock hazard elimination or lightning protection, only lower frequencies are of interest and this "equipotential surface" concept must only apply for local areas. For EMP, on the other hand, where distributed systems are of concern, this concept must apply over large geographical areas and over a broad frequency spectrum.

Another factor is the connection from the system structural members to the earth. The idea is to couple the electromagnetic energy into the earth and absorb it in the earth. The impedance to earth is a function of the connection from the system to the groundwell, and the impedance of the groundwell to the earth. To minimize voltage buildup due to EMP or other transients, these impedances must be low at all frequencies of interest. All conductors have associated with them in inductance which is a function of the geometry of the conductor. The inductance term produces an IZ_c drop (conductor impedance), in addition to the IZ_g (ground impedance). The groundwell impedance can be reduced by using large contact area ground rods or wells and improving the conductivity of the earth in the immediate vicinity by "salting" (ionic salts). Further, most of the EM energy in a current carrying conductor exists in the field external to the conductor. The impedance mismatch results in the field stored energy being reflected and reverberating on the conductors.

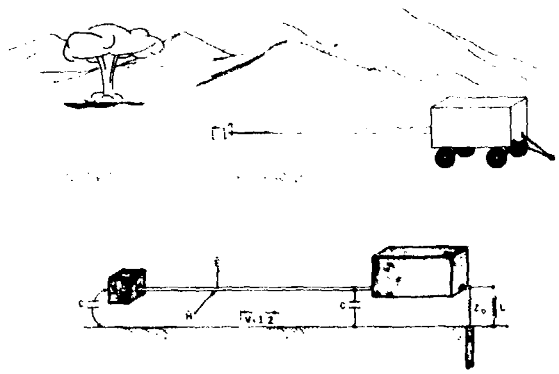
Ground Resistance and Transient Impedance



Most of the Transient Field Energy is Outside the Physical Conductor.

One demonstrated use of controlled grounds was found during atmospheric testing. Here, long duration reverberation currents on the exterior of the cables were present which often interfered with shock wave measurements which occur a few milliseconds to seconds after the blast.

By viewing the long cable run as a transmission line and terminating it in its Z_0, much of the reverberation could be suppressed.



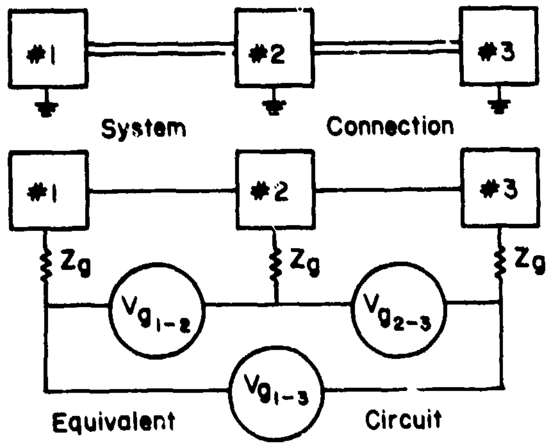
This EMP cable ringing suppression calls for a resistance (10-200 Ω) in series with the ground lead. This is compatible with other exterior grounding requirements. Most grounding systems require a low (<< 1 ohm) impedance earth connection. In power systems, impedances of 10-200 ohms would significantly reduce fault currents possibly inhibiting the operation of protective elements. Further, the fault currents would result in a large voltage appearing between the equipment and earth creating a potential shock hazard to personnel. This latter problem would also exist for lightning grounds. Even communication system's antennas require low impedance ground connections (earth counterpoise) for good radiation efficiency.

For power protection, the use of a choke shunt with low 60 Hz Z, but high EMP Z, can be a compromise for lower power users.

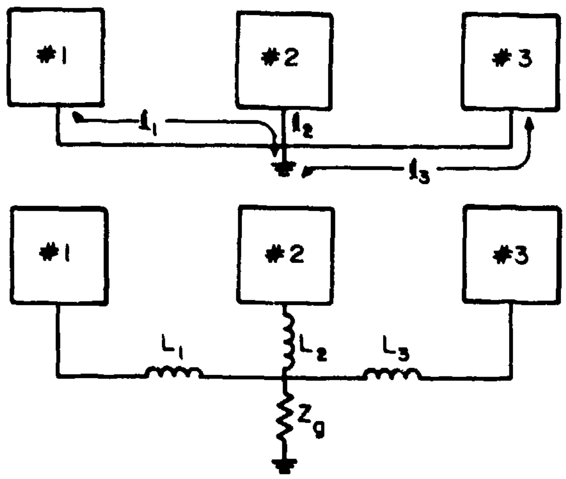
Therefore, good earth connection (i.e., low Z) at all frequencies of interest (EMP or other transients) will provide local protection of personnel from shock hazard, equipment protection in terms of insulation failure, and fault current protection.

Another factor to be considered in designing a grounding system from an EMP viewpoint is EM energy coupling into signal circuitry via the ground system. Two grounding systems philosophies have been expounded in the literature: (1) the multipoint grounds, and (2) the single point grounds. In either case, EMP can induce substantial amounts of current flow onto the shields of various portions of the system and the interconnecting cables or conductors.

In the case of multiple point grounding, ground loops which can couple EM energy through Faraday induction are formed. The voltage induced in these ground loops is equal to the time rate of change of the magnetic field and the area of the loop ($V_g = \dot{B}A$). While such a grounding scheme meets all the requirements for personnel safety and equipment protection, it does not meet the EMP requirements. The voltage induced in these ground loops appears as an offset voltage for signal circuits that utilize system ground as a reference. This can result in upset, and in some cases, damage to sensitive circuits and components.



The concept of a single point ground is to tie all system elements to a common central ground point, thus eliminating the ground loops. These "straight" cable runs can also pick up EM energy. Since the earth is a finite conductor, an 'E' field exists along the ground leads which induces a potential gradient along these leads. If these ground leads are long (>100 meters or so), a local shock hazard exists. Further, since the ground leads have inductance, a high surge impedance can exist which would cause problems in lightning protection and fault current protection. The voltage induced in the ground leads can be calculated



Equivalent Circuit

And Is Given By: $V_{1-G} = \int_0^l E_g \, dl$

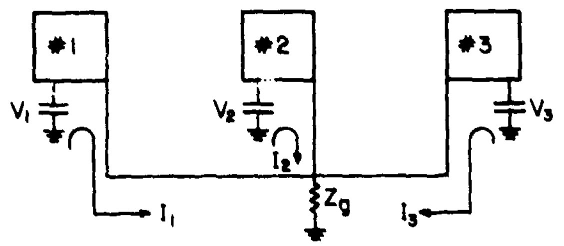
Note If $l_1 = l_2 = l_3$

$V_{1-G} \approx V_{2-G} \approx V_{3-G}$

The problem is further complicated since each element of the system (#1, #2, #3) will have a stray capacitance to ground.

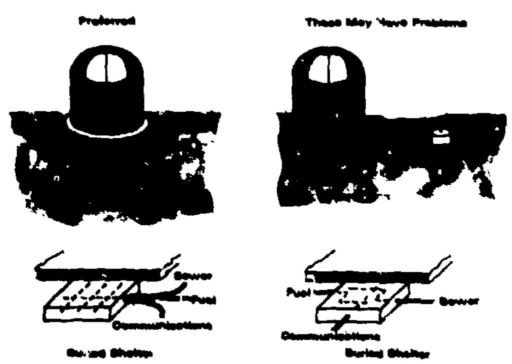
In this case

$$V_1 - V_2 = \frac{1}{C_1}\int_0^t I_1\,dt - \frac{1}{C_2}\int_0^t I_2\,dt\text{, etc.}$$



PRACTICAL SINGLE POINT GROUND

In summary, exterior grounds can be a source of EMP pickup (current collection and/or field enhancement). Neither the single point ground nor the multipoint ground provide an absolute solution. The best choice, in many cases, is to maintain a single point concept as closely as possible utilizing a "tree" concept where the branches to various system elements are of equal length since this is the simplest ground arrangement to achieve control of the collected ground currents. In such a ground system, the power, lightning, and safety grounds should be made as short as possible to minimize EMP pickup. In those instances where these ground connections were prohibitively long, multipoint grounds will have to be utilized and the signal circuits isolated from the induced ground potentials by other means.



Interior/Reference Grounds

The purpose of interior grounds is to provide an equipotential connection. This connection is to eliminate shock hazards between equipment frames, cabinets, etc., and the surrounding structure (van, building, etc.). Further, it is extensively utilized as a reference node for signal processing.

Basically, as in the case of exterior grounds, there are two grounding concepts: the multipoint ground, and the single point ground. Either of these concepts is usable if certain conditions are met.

For systems hardened against EMP, the structure housing the system is a shielding structure. As discussed earlier, the shielding effectiveness for plane waves is relatively easy to obtain if apertures and penetrations of the shield are carefully controlled. The shielding effectiveness for low frequency electric ($\vec{E}$) fields is also easy to obtain, requiring only a good conducting surface. Shielding of low frequency magnetic fields ($\vec{B}$ and $\dot{\vec{B}}$) is another question. To eliminate low frequency magnetic fields requires very high permeability materials and thick walls. This is usually not practical. The result is the coupling effects due to the B and $\dot{B}$ fields dominate inside any shielded enclosure.

Further Facts

Almost Any Metal Envelope is

Very Good E Shield
Fair $\dot{E}$ Shield
Poor $\dot{B}$ Shield
Bad B Shield

With a Good Shield,
$\dot{B}$
B

Coupling Effects Dominate

There is a wide variety of geometrical arrangements for "connecting to ground." Here we identify several of the accepted connection geometries:

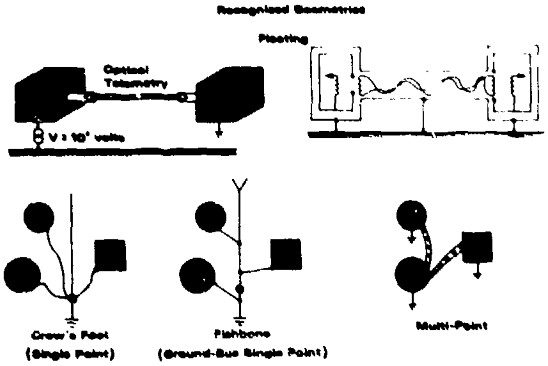
"Crow's Foot" or single-point -- probably the wisest choice in a "bleak" situation, since it minimizes coupling in the ground connections proper.

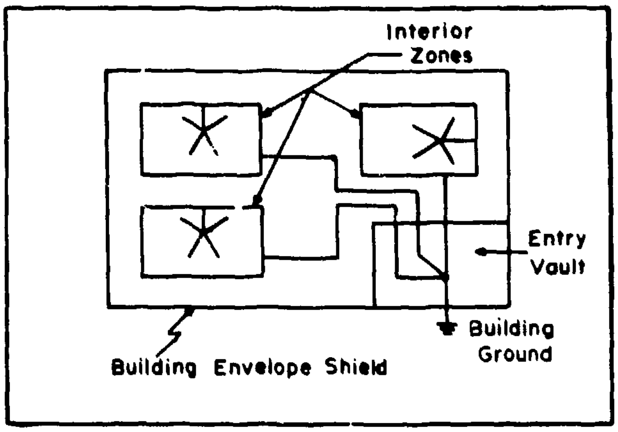
Fishbone -- the lower level (higher sensitivity) circuits should usually be at the "far end," where "ground currents" are lower.

Multipoint -- note the opportunities for ground loops and common impedance IZ_{common} voltage rises.

Floating Grounds -- often employed where a single point ground is impractical and where a multipoint system could cause trouble. Here, each subsystem case assumes its own potential without ill effects, provided that good isolation (common-mode rejection) is realized.

In large facilities, since the classical "star" or "crowsfoot" is not practical for the reasons discussed, it is often replaced by a "TREE" system. The "TREE" system is a combination of Faraday shielded zones, each utilizing a classical "star" system ground, for signal reference and local hazard, with the local Faraday shields grounded to the main structure ground. These zone grounds are for safety ground only and signal grounds are isolated by the zonal shield connections. These grounds may be obtained by ground buses or directly to the conducting structure depending on the facility geometry. Signal connections between zones must utilize wired return and not provide additional ground paths between zonal shields.





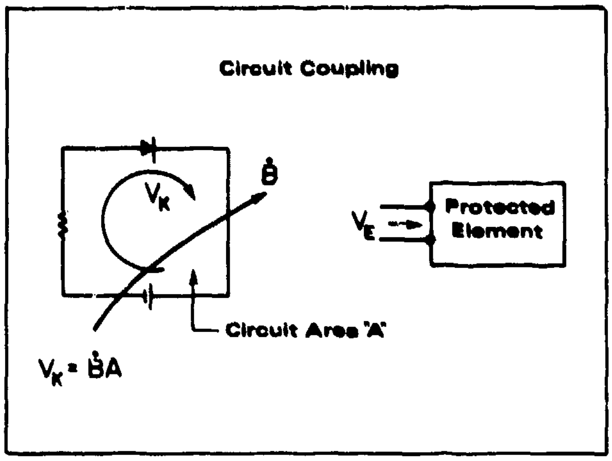
Many of these configurations work very well if the ground lead length is short compared to the shortest wavelength of the penetrating fields, insofar as coupling energy into the system via the ground leads. The multipoint ground also provides for loop coupling. An additional problem that can exist with the multipoint ground is that each element of the system is not tied to an equipotential surface, that is, if the surface is the enclosing structure, it may not be at a single potential due to bad seams, etc.

All ground leads also exhibit an impedance, usually inductive reactance, which varies with frequency. This restricts the use of the single point ground as a reference ground, especially in large enclosures where these ground leads may be quite long. This would introduce reference level voltage offsets between various portions of the system. To eliminate these reference level voltage offsets, isolation of the signal circuits can be utilized as shown by the now-conducting transmission circuit or the balanced circuit (floating ground system).

Circuit Considerations - Circuit Coupling

There are three predominant ways in which EMP energy gets into circuits: (1) B -dot doupling, (2) terminal injection, and (3) direct injection. The effect of $\dot{B}$ coupling can be analyzed as equivalent to distributed circuital voltage input. Terminal injection appears as extraneous voltage or current pulses at the peripheral equipment terminals. Direct injection results via a mutual coupling element, such as a "multipoint common ground."

Circuit EMP can also be directly induced by E-field coupling or by EM radiation exposure (antenna effects), but these are usually not important in enclosed, well-shielded systems. One should also distinguish those situations in which EM or E-field coupling induces currents (or voltage) which subsequently couple by B-dot or conduction effects (internal zone conversion).

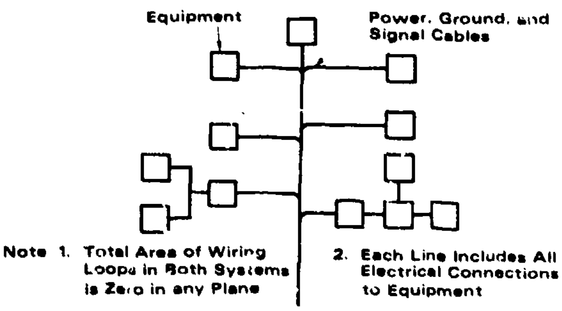


Perhaps the single most significant concept in curcuit considerations is to recognize that the same fundamental rules of coupling and response apply to both large and small conducting circuits.

Thus, large cabling layouts and small printed circuit packages are basically amenable to the same analytic approaches. Earlier we saw this to be the case in B-dot coupling considerations. Only the dominant frequency tends to be determined by circuit dimensions and reactances. Further, the same types of fixes apply to both large and small circuits, although the implementation of these fixes may differ.

Control of EMP coupling due to the B-dot mechanism can be achieved, in large part, by control of the circuit layout. The principle idea is to eliminate coupling loops through a "TREE" wiring system. The concept is to establish a limited number of interconnection points of compact and controlled geometry.

Circuit Control



Wiring to Reduce EMP Susceptibility

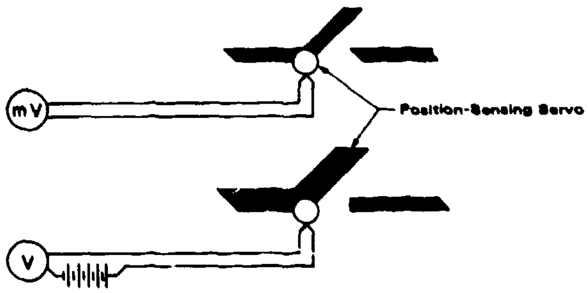
"TREE" Wiring System

Circuit Configuration

It is necessary to provide communication between various systems and subsystems. Several approaches were introduced during the discussion of the systems aspects of EMP hardening. These alternative approaches will be expanded upon in this section.

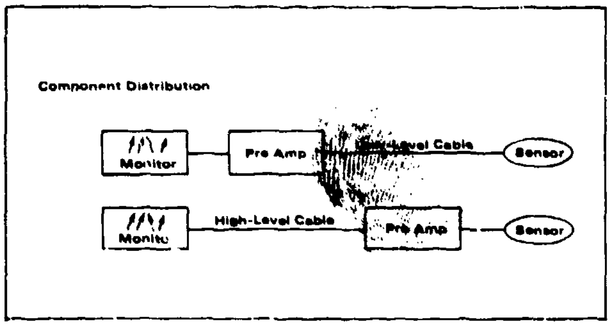
One approach to reduce the protection requirements of the cabling system, and consequently the cost, is to configure the system to utilize high signal levels in the longer cable runs. Operation at higher signal levels results in increasing the signal to noise (unwanted energy) ratio. Since the system operates at higher voltage levels, it implies the use of inherently harder components. The implications with regard to circuit upset are obvious.

High Operating Levels



Component Distribution

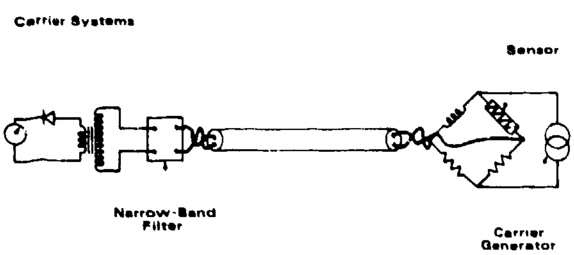
Closely related to operating level is the distribution of components. A simple example appears here. Since the EMP coupling is primarily due to the cable, the system utilizing the high signal level, preamp at the sensor, will be inherently hardest. Note too, that the choice reflects on the character of the terminal circuits as well. In the preferred configuration, the preamp outputs and monitor inputs will tend to be "harder" simply because they must operate at higher signal levels in their own operation.



Another hardening technique is to go to carrier frequencies. Many sensing and control situations lend themselves to this by relatively inexpensive terminal hardware -- provided this choice is made soon enough. Due to the added cost usually associated with this approach, its use in EMI has been limited. However, additional criterion of EMP hardening could easily tip the scales in favor of carrier systems, rather than dc or low-level, self-generating circuits.

Carrier systems have the advantages of permitting floating balanced conductors, narrow band-pass filtering, transformer isolation, easier nullification, and much more.

A Simple Circuit Example



The carrier frequency chosen to implement this approach should be well above the highest frequency of the incident EMP spectrum. This provides for the use of high pass filters with good roll off characteristics. Spurious responses below the nominal cutoff frequency of the filter must be carefully controlled.

Carrier systems, of a more sophisticated nature, may employ dielectric waveguide operating in the microwave spectrum or optical data links.

Generally, the EMP susceptibility of a cable system is related more to the sensitivities of the terminal elements and circuits than to "breakdown" or "burnout" limits in the cable itself. This points at once towards terminal protection and we will say much more about this, but a circuit designer may be able to improve matters by keeping EMP in mind when he considers the terminal element design and the circuit routing through the cable system.

Cabling Design - Cable Types

In many cases, hard wire intersystem connections are the only alternative due to cost constraints, or the type of information to be transmitted. Therefore, it is appropriate to discuss the various aspects of cable design.

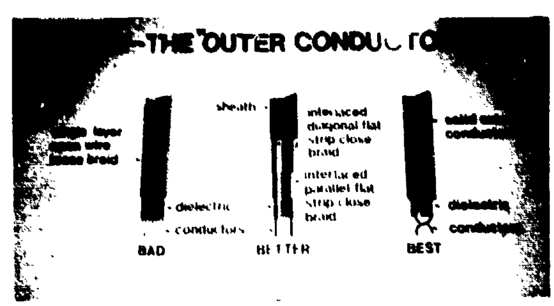
When one examines the possible permutations of cable component choices, it is obvious that no detailed case-by-case evaluation is possible. Rather, we can point to certain preferable choices in each category of component.

These can be broken up into two broad categories: those aspects which influence the control of the effect of external environment, and those which control the inner cable environment, circuit intercoupling, and so forth.

Cable Types

- Outer jacket : none, insulating
- Outer conductor : none, solid, braided, wound strip
- Inner conductors: single coax or twin biax separate circuit types inner coaxs, twisted pairs, multi-conductor
- Inner shields : none, braided, thin strip insulated, non-insulated

Consider control of the effect of the external environment, that is, the reduction of the energy which diffuses into the cable. This is accomplished by providing an overall shield on the cable. These shields may vary from a single layer braid type covering, to a relatively thick solid outer conductor.

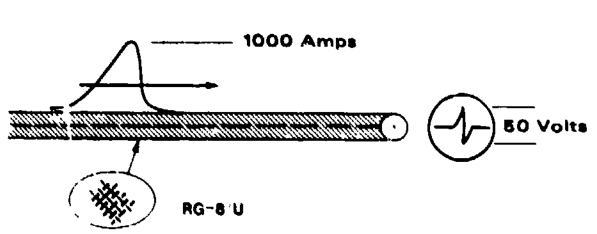


Transfer impedance was discussed previously in the analysis section, but, for completeness, we will consider it again in context of cable design.

If the wires of an unshielded bundle also interweave, then the "surface current" due to EMP exposure gets transferred inside and all of the circuits share in its pickup. A braided shield also behaves somewhat like this. Besides having holes for field leakage, the braid wires transpose. The wire-to-wire contact is not very effective and much of the surface current gets inside, to radiate into the internal cable zone.

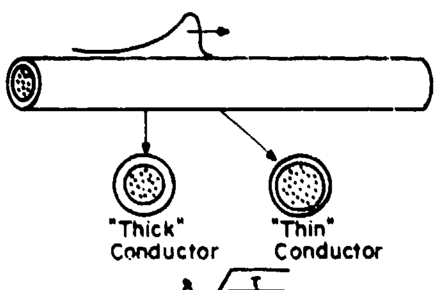
As an example, an RG-8/U cable has a 10^{-2} current transfer ratio or transfer impedance, for sub-microsecond pulses. Thus, at the end of a 50 meter length exposed to a field pulse which induces 10^{+3} amps peak current, there will be about a 50 volt signal on the inside.

Braid Transparency



Use of a solid outer conductor results in a much lower value of transfer impedance at the higher frequencies (above approximately 1 MHz). This would result in much lower coupling to the interior cables (wires). The thickness of this outer conductor, however, is very important at the low frequencies. The outer conductor must be several skin depths thick at the frequency of interest to achieve 60-80 dB of shielding effectiveness. This can be accomplished by a choice of an outer conductor permeability and thickness.

Why is the "Exterior" Important?

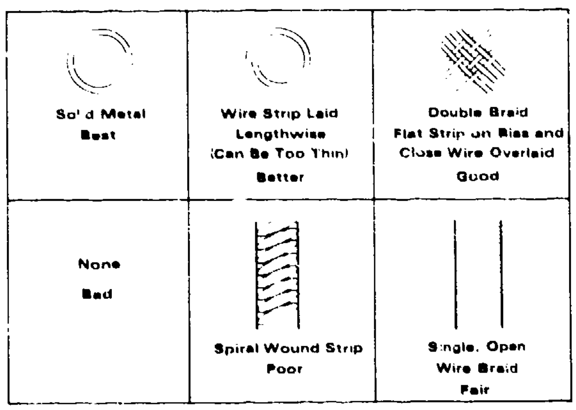


For buried systems, a technique which is often desirable is to use a conducting asphalt as the outer protective jacket over the shield. This conducting jacket helps to prevent build-up of the current wave. Energy induced away from the terminal end of the cable is damped in this manner.

Here is an additional illustration of "good-bad" choices from the EMP standpoint. Like normal enclosure shielding, the thickness is not very critical. One finds that the emphasis is rather on shielding "tightness." Again, we want no "cracks" if at all possible.

One form of "nontightness" is that represented by a spiral-wound strip -- even the double-layer variety. The trouble here is that each turn is actually a turn in a loosely coupled continuous mutual inductance. The contact resistance along the overlaps is too high to avoid some voltage buildup per turn. This type of shield acts as a relatively good coupler between external environment and internal wires.

Outer Shield Construction



Conduit, when properly installed by threaded connection or welded joints, behaves very much like an additional solid-metal outer conductor. When needed for other reasons, such as blast resistance or code requirements, it is inexpensive in terms of added EMP cost.

These problems can be compounded by loose material specifications, poor quality control, and lax acceptance criteria. Often, cables are delivered with the outer shield badly oxidized or even corroded. If the construction of the shield is such that good shielding depends on good internal contact, it will not be acceptable.

In most cases, the outer conductor (shield) is covered by an outer protective jacket. Lead sheathing or neoprene jackets seem to provide cable lifetimes that are commensurate with normal system life. There is some evidence that the insulated outer jacket reduced the coupling for cables in direct field exposure, as compared with no insulation. The conductive outer jacket, as mentioned previously, can provide propagation damping for conducted components of current induced from a distant point. From a cost standpoint, neoprene is the favored approach.

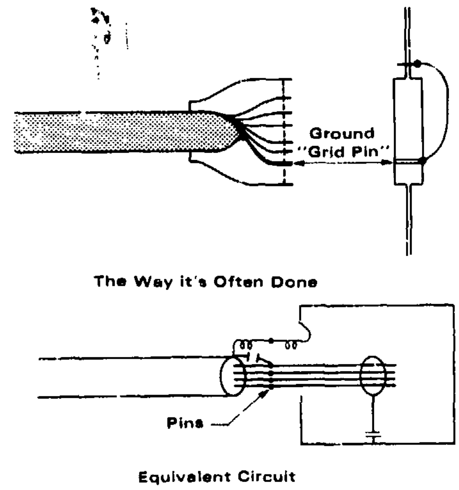
Another way of achieving the propagation damping effect, which has been used in some nuclear test programs, is to use conducting baffles. The idea is to radically change the impedance of a cable over a ground plane resulting in reflecting the energy at the impedance discontinuity. An example of this effect is the entry way for electric power where the overhead transmission line (wire over a ground plane) enters a facility via conduit, the transition being made at the power pole. The impedance discontinuity reflects an appreciable amount of energy.

Coupling of the propagating energy into a lossy material can also be used as a means of propagating damping. This has been implemented in the form of ferrite beads or cylinders around the outer conductor or, in some cases, single wires (either electrical or mechanical use). It essentially provides a lossy filter.

Termination of the Outer Conductor

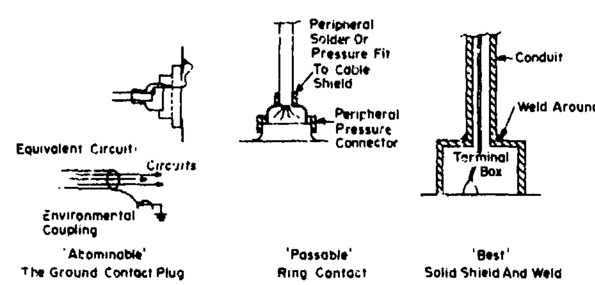
To preserve the protection provided by a good cable shield, it must be properly terminated at the entry into a shielded enclosure or equipment cabinet. Past, and in many cases present, practice is to provide the shield connection via one or more pins of a multiconductor connector. The reason for this was, in many instances, the connector shell was nonconducting. Most connectors are made of aluminum which, in many cases, was annodized for corrosion protection. The effect is to insert an inductance in series with the outer conductor which produces a series voltage drop and couples capacitively to the other conductors. On a transfer impedance basis, this type of termination will exhibit a transfer impedance ranging from 50 mΩ to 10 Ω over the frequency range of 0.1 MHz to 25 MHz.

Cable Terminal Treatment



The cable shield (outer conductor) should make good peripheral contact to the connector shell, and hence, the shell to equipment enclosure to eliminate this problem. Providing a pressure fit between the connector shell and mating the connector to the enclosure via the appropriate bulkhead connector, the transfer impedance was reduced to 5 mΩ at 0.1 MHz, 70 mΩ at 25 MHz, and 300 mΩ at 100 MHz. For comparison purposes, the cable normally used with the connector tested has a transfer impedance of 11.4 mΩ to >1.0 Ω over the frequency range of 0.1 MHz to 100 MHz.

PLUGS AND SOCKETS



This indicates that peripheral bonding is the best solution. Soldering or welding the outer shield directly to the enclosure wall would further reduce the connector transfer impedance. This is only cost effective and worthwhile for cables with better shielding effectiveness than the previous case.

As we hinted before, when properly installed, conduit is easily the best "outer shield" for cable systems. The principal problems arise at segment contacts, of course, Again, cleanliness and careful assembly are essential. Rusted and corroded threads and bushings will introduce series impedances along the conduit's length. Welded joints are best, but expensive.

Welding is almost the only dependable way to deal with the conduit terminals. Normal clamp rings make contact at only a few points, at best. All too often, there are some loose ones left behind. The conduit ends are particularly sensitive system points, because here the exterior transmission impedance changes. Pulse currents flowing on the outer surface must be redistributed onto the equipment shield.

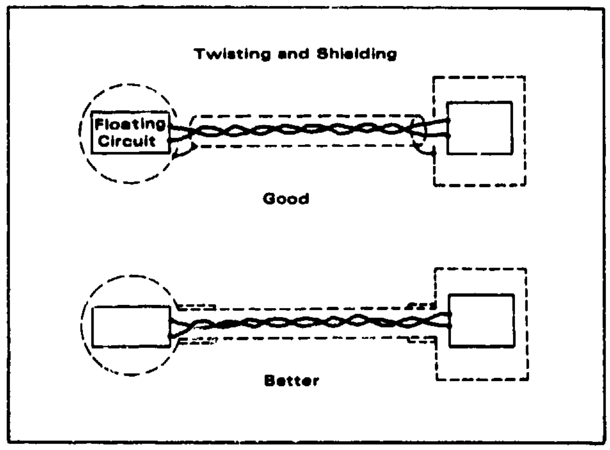
If relative movement between exterior conduits and shielded buildings are expected (due to shocks or earth movement), the use of bellows, convoluted sections, or multiple knitted socks should be considered.

Internal Conductors

There are two basic cable constructions which are of interest: (1) the coaxial cable, and (2) the multiconductor paired cables. Coaxial cables use the shield (outer conductor) as the signal return. Pickup of extraneous signals is primarily due to the transfer impedance mechanism for well constructed cables. Cable eccentricity can also result in coupling of magnetic fields but is less prominent. Good coaxial cable design, from an EMP coupling viewpoint, is primarily an effective shield design. As discussed in the previous subsection, the best shielding effectiveness can be obtained with a solid outer conductor. If cable flexibility is required, double braided shields can provide satisfactory performance if properly designed.

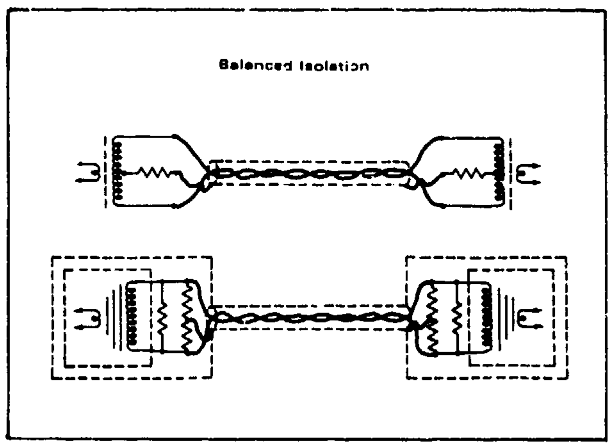
For medium-level sensors and medium-bandwidth circuits, adherence to zoning and circuit reference criteria may suffice internally. This dictates provision of separate return reference wires (independent 2-wire circuits) which are fabricated as a twisted and shielded pair. Braided shielding is often sufficient here. Broadly, these practices are generally commensurate with requirements for intercircuit isolation ("cross-talk").

The single reference connection bond is not as desirable as a continuous metal backshell connector which grounds the cable shield to the subsystem shield without introducing apertures or high-impedance connections.



As indicated before, if a system lends itself to carrier techniques, then we can gain 30-40 dB in protection value by using balanced cable circuits with terminal isolation transformers. Conversely, the requirement on the built-in cable shielding may be that much less severe.

The advantage of the "carrier" approach resides in the ease of obtaining balanced isolation transformers that operate over a "limited bandwidth." Extra filtering may also be required to assure that the bandwidth is restricted to the "limited bandwidth" of the transformer.



Multi-layered shields is a good approach for sensitive circuits. However, if the shields are not properly terminated, the coupled energy will reflect from the terminal ends of the cable, resulting in additional penetration of the cable. Two approaches are shown: (1) terminating the inner shield in dissipative loads, and (2) isolation of the shields. To decide the means of terminating the shields, it is good practice to consider the multilayer shields and common-mode circuits as independent transmission lines and terminate them accordingly.

As discussed in Section IV, the lowest energy coupling is achieved for a non-carrier system by the shielded twisted pair. If care is excercised during cable construction, good common-mode and differential mode protection can be achieved.

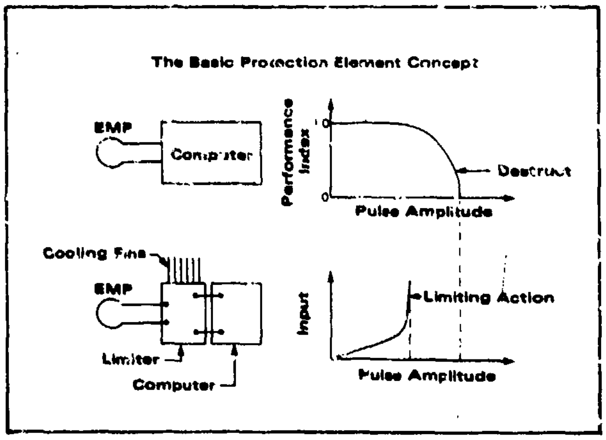
Some "connections," such as power supply and control circuits, must be dc or non-carrier ac. If these can be treated by means of terminal filters and by high-level operation, then (together with carrier signal techniques) the EMP requirement on the total cable package can be a minimal one.

Protective Devices and Techniques

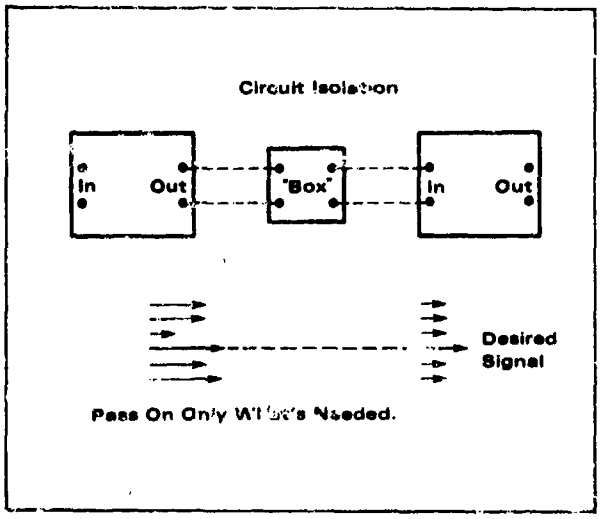
The discussion to this point has been concerned with system protection by keeping the energy out of the system; that is, through the techniques of reducing the coupling and shielding. These approaches are necessary for hardening a system; however, they only reduce the amount of energy entering the system. Completely eliminating coupling to or ultimate shielding of a system is not viable.

The result is that some energy will enter the system which may effect the performance of the system. The use of protective devices at the equipment terminals is another means of hardening at the subsystem/circuit level.

Protective devices are used to divert or dissipate the undesired electrical surge energy. An efficient protection element has a performance characteristic which is inversely proportional to the susceptibility of the hardware or function which it is intended to protect.



The protective devices fall into two major categories: (1) amplitude limiting devices which clamp the voltage or current to the desired level, and (2) spectral limiting devices which remove energy in certain frequency bands. These approaches may be used singularly or in combination (hybrid devices) depending on the system and/or protection requirements. The idea is to pass only those signals, in terms of amplitude or frequency, that are required for normal operation of the circuit.



Spectral Limiting Devices

These devices are utilized to suppress certain frequency components of the EMP induced signal. In considering spectral limiting devices, the waveform of the induced signal must be considered. Normally, the induced EMP signal will appear as a damped sinusoid whose dominant frequency is determined by the point of entry and the characteristics of the coupling path (Section IV).

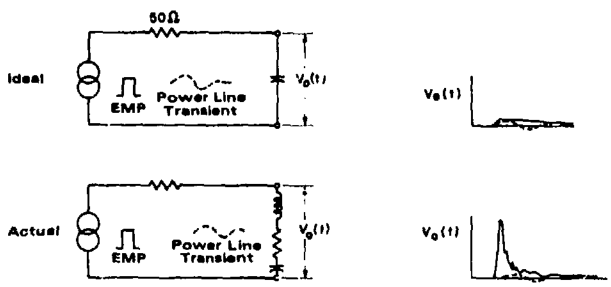
Included among spectral limiting devices are capacitors, inductors, ferrite beads, transformers, bifilar chokes, and combinations of R-L-C type filters. Spectral limiting devices can serve two basic purposes in regard to EMP suppression: (1) they can suppress spurious frequencies, and (2) they can reduce the wavefront slope which would allow a slow spark gap to respond. One factor which must be considered in using spectral limiting devices is they only suppress frequencies outside their pass band and thus provide little or no protection if the EMP induced signal falls within that frequency range.

Capacitors

Capacitors are often used to bypass the switching transients and lightning surges on power lines. The "line" appears as a 50 to 600 Ω surge impedance, and if bypassed by an ideal capacitor, the spike is integrated.

In practice, this will not work for early-time EMP induced signals because capacitors have series impedance R and L which are significant EMP-wise for the first 100 nanoseconds.

Examples of Filter Responses



Capacitors are Often Used to Suppress Power Line Switching Transients

This shows test results for the circuit just considered. Note that a 30-50 nanosecond spike is passed by the "filter" and that lead dressing affects the response downstream of the capacitor at late-times.

A Generator into 50 Ω Load

B 50 Ω load By-Passed by 1µfd with Typical Lead Dressing

C 50 Ω Load By-Passed by 1µfd with Very Short Lead Dressing

Scale 1 unit = 10 nanoseconds

Although capacitors do not provide the required protection alone, they are still very useful to eliminate "noise" caused by surge arrester switching on the leads exiting entry vaults or enclosures.

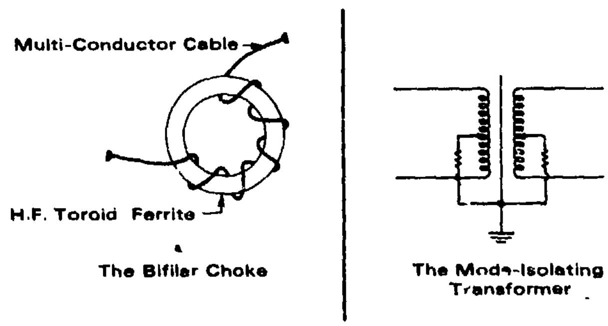
Inductive Devices

Two inductive devices are particularly useful in common-mode suppression, as may be required on cable connections between equipment in two different EM zones. Both are extensively used in electronic instrumentation.

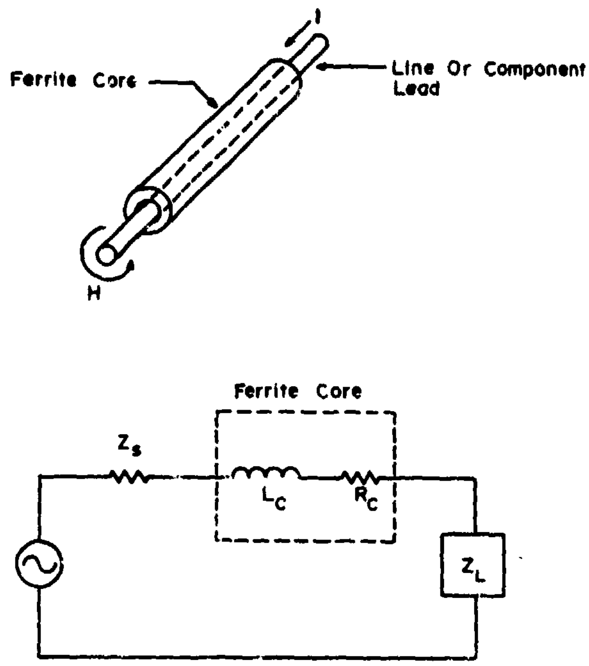
In a bifilar choke, the push-pull or desired circuit paths carried by the multiconductor cable are only weakly coupled to the core, whereas the common-mode carried by the whole bundle is strongly coupled. Hence, the common-mode is strongly discriminated against. The required series inductance depends on the desired attenuation and on the predominant frequency content.

The balanced mode transformer similarly discriminates against common-mode energy, but is limited to ac/HF application.

Passive Devices - Inductive



Ferrite beads can also be used to suppress unwanted energy. These are ferrite toroids or cylinders slipped on individual wires. These devices are dissipative and nonlinear. The nonlinear properties vary with frequency. Ferrite beads suppress frequencies above 1 MHz whereas ferrite chokes may be used at frequencies as low as 20 kHz with special design. One disadvantage of these devices is that due to dc currents, they saturate quite easily. Saturation can occur in particular types for currents as low as ten (10) milliamperes.



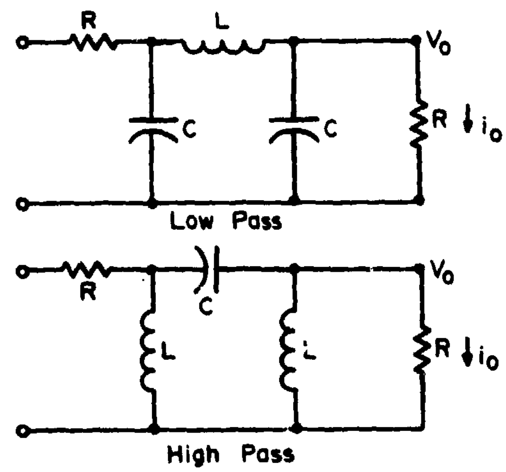
FERRITE CORE AND EQUIVALENT CIRCUIT

Filters

The most common passive, lumped-element device is the terminal filter. It is basically a black-box with input and output connections for insertion into an otherwise continuous two-wire circuit. Its insertion loss is chosen for minimum attenuation in the frequency domain of normal circuit operation and maximum attenuation in the domain of maximum "noise" (i.e., induced EMP signal) content.

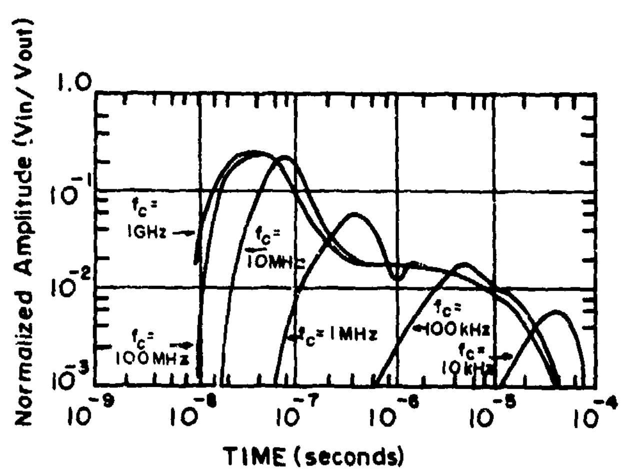
The intercepted energy has to be diverted or absorbed. It may be reflected back into the input system, increasing the EMP level there. A better alternative would be to dissipate the energy as heat in an internal filter resistance. This implies a preference for "lossy" filters.

The effectiveness of the filter also depends on the out-of-band responses of the filter. Since the EMP induced energy contains a wide spectrum of frequencies, even the damped sinusoid, the filter must not have any spurious responses over the interfering signal spectrum. Further, if filters alone are used, the components used in the filter must be capable of withstanding the induced voltages and currents without failure.



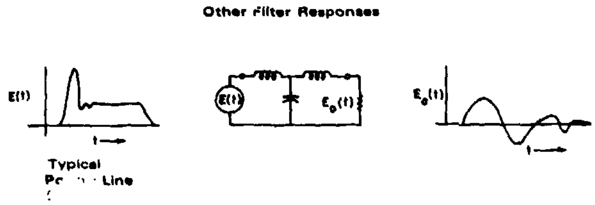
Here is a typical working example of a filter analysis for a very severe exposure situation. One might experience something as bad as this for a completely "naked" megawatt level power line. The figure shows the effect of a simple, three-element filter for various design parameters.

Note the logarithmic scale compression. Obviously, the linearized pulse will be much "more peaked" in appearance.



It should be noted that the lower the cutoff frequency of a low pass filter, the more the slope of the wavefront is reduced.

Another example shows the rejection for a typical low pass filter. Shown at the left is a typical power-line EMP surge. Assuming the filter cuts off at about 15,000 Hz, it can pass about 1/5 of the applied energy for certain expected applied waveshapes.



Amplitude Limiting Devices

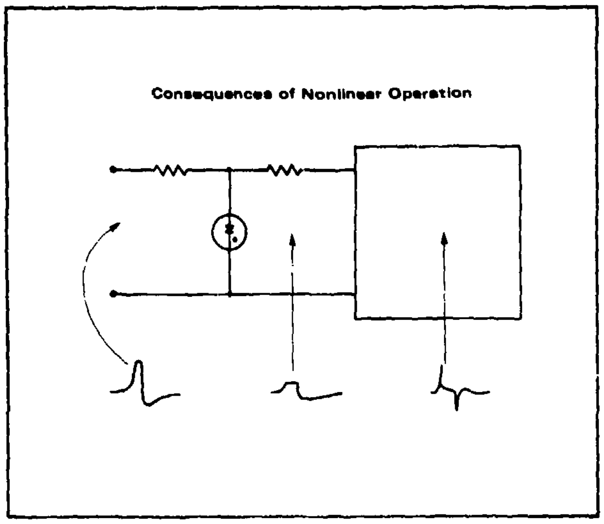
Amplitude limiting devices are usually nonlinear in nature. Included among these devices are spark gaps, gas and semiconductor diodes, soft limiters such as thyrites and varistors, and mechanical or thermal devices such as fuses, circuit breakers, and relays.

Types of Non-linear Devices

- Spark gaps
- Gas diodes (cold, hot)
- Zener diodes
- Silicon diodes (w/bias)
- Thyrite (sic)
- Fast relays
- Hybrids
- Crowbar circuits

Nonlinear devices can cause some problems. We already indicated under the heading "Filters" that the EMP energy has to go somewhere. This remains true for active elements as well. Furthermore, the switching operation itself can be a source of unwanted EM energy (e.g., RFI) interfering with sensitive downstream components. This is particularly true if the associated circuits contain significant EM energy in normal operation. When the device switches, it must inevitably cause some change in effective circuit impedance and hence in operative current distribution.

In addition, the switching function may generate a spurious pulse in the circuit itself. This is particularly possible if the switching occurs on a time scale short compared to that of the normal operational signals in the system, e.g., on the fast-rise "front" of an induced EMP signal. This is one of the strongest reasons for using "hybrid" (i.e., limiter/filter combination) lumped elements.

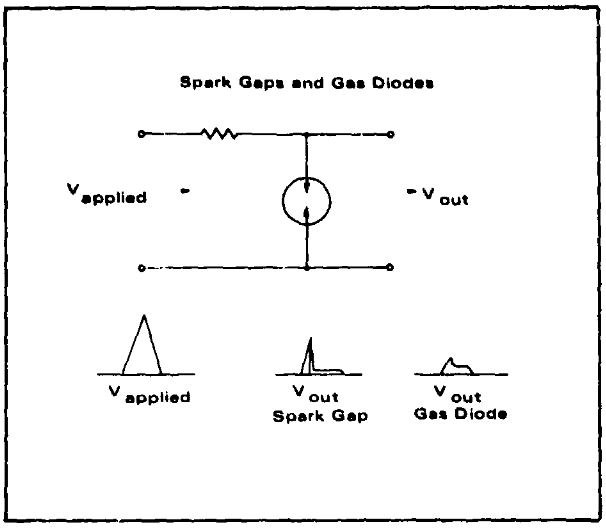


Spark Gaps and Gas Diodes

A spark gap is a voltage-threshold switch with two or more electrodes separated by a dielectric gas. Spark gaps depend on initiating conductive breakdown in a gas. When this breakdown occurs, the device switches from a very low to a very high conduction state.

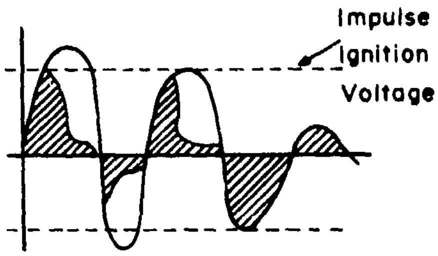
Spark gap type devices have the advantages of being bipolar in operation and can handle extremely large currents (thousand of amperes). They are available with static initiation voltages of 60 to 30,000 volts. The initiation voltage depends on the gas medium, gap spacing and gas pressure. Arc initiation requires some free electrons and, therefore, there is a minimum for a given gas. The lower voltage devices often use radioactive doping to increase the number of free electrons and thus lower the initiation voltage.

Since these devices require mobilization of the free electrons under the influence of the applied field and generation of secondary electrons due to collisions, it takes a finite time for the arc to occur. Because of this time delay, these devices will fire at the static breakdown voltage for slowly rising pulses, but require higher voltages for rapidly rising pulses. This impulse ratio may be anywhere from a few percent of the static voltage to several times the static voltage depending on the steepness of the wavefront slope. For example, a 500 volt gap may have an impulse voltage of 9500 volts for a pulse having a rate-of-rise of 5 kV/ns.



The extinguishing potential must also be considered for spark gap type devices. The voltage at which the device extinguishes is a function of the current through the gap; the higher the current, the lower the extinguishing voltage. This can be a serious problem in dc circuits where the follow current from the source is sufficient to keep the gap ionized. In ac circuits, the gap will extinguish as the signal passes through zero. Special arresters (expulsive arresters) use magnetic fields or gas emissions to extinguish the arc. Resistance in series with the gap (fixed resistors or varistors) can be used to limit the surge current, but this results in higher terminal voltage.

The operation of a typical gap arrestor for a damped sinusoid exciting voltage is shown. The total energy passed is indicated by the shaded area. It should be noted that if the signal amplitude is insufficient to ignite the gap, more energy can be passed to the protected circuit.



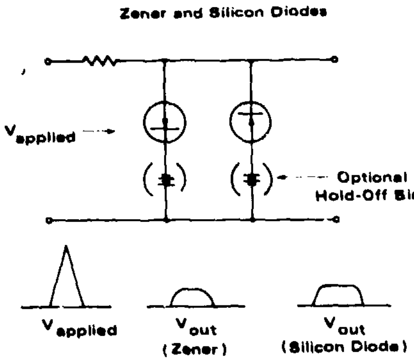
Zener and Silicon Diodes

These are generally smaller, lower power devices. They operate effectively in the voltage-current range of solid state circuitry, so that they are extensively used for such circuit protection. They are voltage-limiting in action (rather than voltage-reducing). Silicon diodes "clip" more effectively -- their "plateau" is flatter. Their operating voltage is generally low -- a few volts -- and the introduction of "hold-off" bias can be an inconvenience. They are generally high-capacity devices, so that there are limits as to the circuit frequency range of applicability. Also, the semiconductor devices have definite limits on the joule energy handling capability (i.e., the protective elements may have low "damage" thresholds).

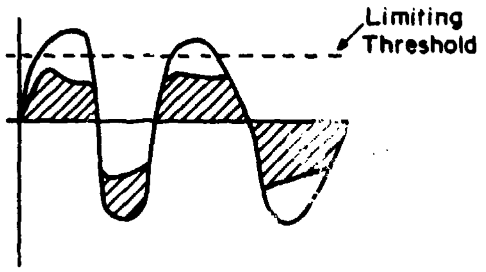
In general, these devices do not exhibit a turn-up (impulse) voltage characteristic. The p-n junctions in these devices will react sufficiently fast to limit transients even with nanosecond rise times. Caution must be exercised in the application of these devices since if very short lead lengths (less than 1/2 inch) are not utilized, the inductance associated with the leads can result in a voltage drop far in excess of the normal junction voltage.

Semiconductor limiters (diodes) are unipolar devices. The limiting is usually provided by the reverse voltage breakdown characteristic. When breakdown occurs, large currents can result causing a possible failure of the semiconductor if care is not exercised. Zener diodes which are often used for this purpose range in clamping voltage from about 2 to 200 volts. Current handling capability may range as high as a few hundred amperes. For very short (100 of nanosecond to microsecond) pulses.

Extinguishing of these devices is no problem since there are no residual electrons to be swept as in the case of spark gap type devices.



The response of a zener diode limiting circuit to a damped sinusoid incident signal shows the hard limiting features of such a device. The case shown is the same incident signal as that shown previously for the spark gap arrester.



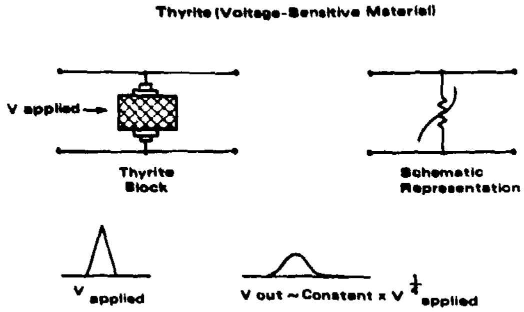
Varistors

A varistor is sometimes referred to as a "soft" limiter. It is a bulk semiconductor material whose resistance is varied with the magnitude, but not polarity, of the voltage applied to it. The voltage/current curve is nonlinear but never negative. Therefore, the voltage drop across the device always increases.

There are two basic types of varistors available: (1) the silicon carbide (SIC) type, and (2) the metal oxide varistor (MOV). For very low values of current (less than approximately one microampere) the device acts as a linear resistor with a resistance of hundreds of megohms. At higher values of current, the voltage current relationship is nonlinear and is given by:

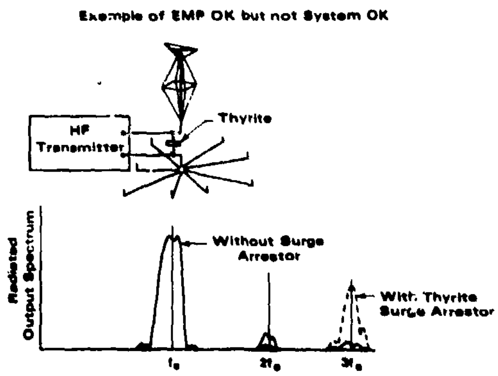
$$I = KV^{\alpha}$$

For silicon carbide varistors α ranges between 2 and 7, whereas for metal oxide types an $\alpha \approx$ -25 is typical.



MOV's are available with clamping thresholds of 40 to 1500 volts, response time in the nanosecond regime, and peak energy handling capability of up to 160 joules. SIC's have considerably higher peak energy handling capability, 270,000 joules, and clamping thresholds of 15 to 10,000 volts.

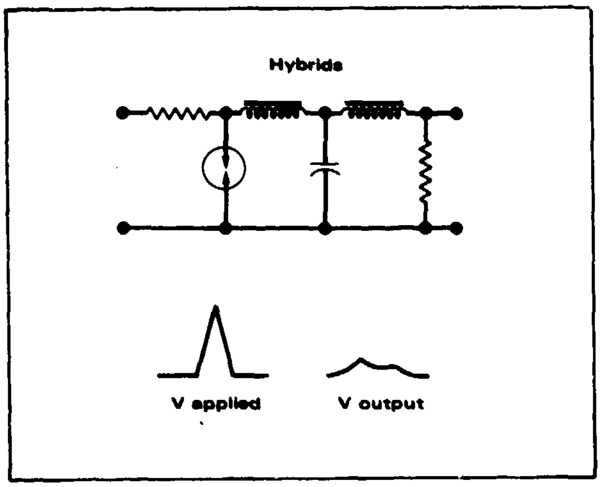
Since these devices are nonlinear resistors, care must be exercised in their application. If they are applied to circuits where normal voltage swings cause a resistance change, the device can generate harmonic and intermodulation type interference which may be objectionable. This could be a problem in RF transmitter output circuits or in power circuits.



Hybrids

A hybrid circuit, that is, one employing a combination of amplitude and spectral limiting devices, is one of the more favored approaches for terminal protection. A spark gap or other high energy amplitude limiting device can be employed to shunt the bulk of the current. The filter (low pass) following the gap reflects the high frequency energy in the spike resulting in a slowed rate of rise of the wavefront. It also reflects the high frequency noise associated with gap firing. If the pulse out of the filter is still too large, a second, low energy arrester such as a zener diode can be used since the energy in the pulse has been reduced to safe levels for the zener.

The series impedance preceding the surge arrester is a necessary component to assure appropriate limiting; in some cases, the surge impedance of the transmission line can suffice.



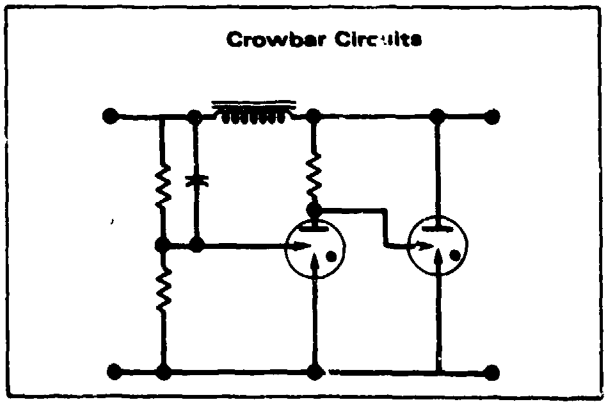
Electromechanical and Thermal Devices

Fast relays, circuit breakers, and fuses have response times that typically are on the order of one (1) to several milliseconds. Their principal value lies in interrupting circuit operation as a result of the current "dumping" of the faster protective devices (arresters). This circuit interruption feature limits the energy which must be dissipated in faster devices and the relays may also be used to initiate restoration to normality from a breakdown condition.

Crowbar Circuits

In these systems, a high-power rating device is operated by a subsidiary sensing/trigger circuit. Thyratrons, ignitrons, spark gaps have been used for the "crowbar." Sensing can come from the circuit itself, from a "threat" sensor, or from an auxiliary breakdown device (e.g., corona optical sensor).

Crowbar circuits are often used to activate the normal system protection interlocks. For example, EMP could "fire" a spark gap or cause an arc-over in the transmitter output. This arc, if not extinguished, could cause excessive plate dissipation in the output tubes. In this case, a thyratron can be "fixed" across the transmitter dc supply to activate the circuit breakers. The thyratron is activated by a corona-sensing photocell near the spark gap or better yet by an impedance sensing circuit.

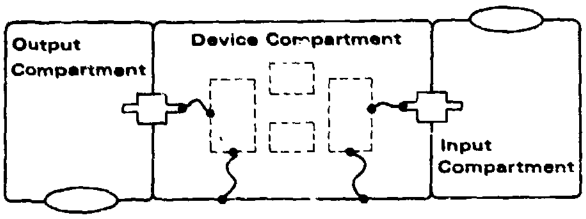


Device Construction and Installation

The construction and installation of a protective device is often as critical as its design. If we think of a filter as a controlled RF barrier, then it is clear that its input and output must be electrically isolated from one another. A good filter (or other device) is usually constructed in three separate electromagnetic sections, as shown here.

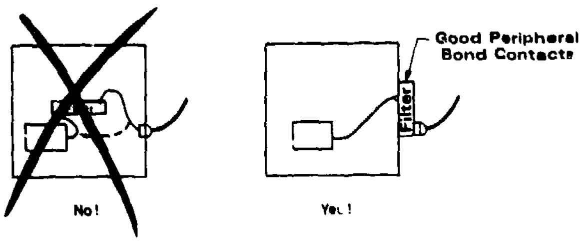
Most frequently, filters and limiters operate "against ground;" that is, the "return" side of the protective element is well bonded internally to the filter case. Good filter design and adjustment takes into account whatever mutual coupling may exist between input and output within the central component compartment. This convention comes from the customary circuit practice of using "case" as the reference node in small and medium size system elements, both for single-ended and balanced systems.

Device Construction



Obviously, the same care in isolation is called for in installation; much of the device's value is lost if the output side can electrically "see" the input side. In the "right way," the filter case must make a tight peripheral contact so there is no "hairline" aperture and so the common reference impedance is nearly zero. This is also important if one is to obtain the benefits of the designer's and manufacturer's ratings.

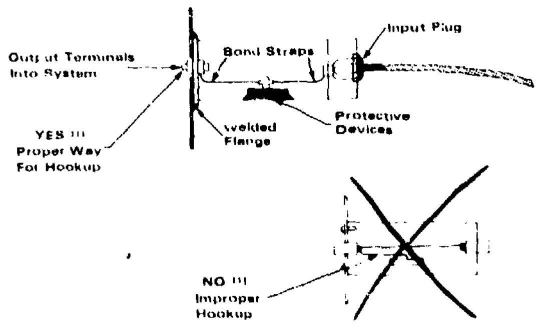
Device Installation



The most serious defect in commercial protection devices is the penchant for using "pigtail" type connections between terminals and the protection element itself. These generally present at least as high an impedance to an "EMP" as does the circuit itself.

Many of these devices are useful for EMP protection if they use low-inductance bond straps and adequate connection contact areas -- especially to the case (or "common reference") side. In most cases, the signal circuit should be taken through the box; the protective device should not simply be shunted at a single terminal point.

Active or Non-Linear Devices



In some installations, particular care has been taken to isolate such entrance protective devices. This is the origin of the "EMP Room," sometimes ostentatiously displayed as the "solution to EMP." In older, "unprotected" systems, one finds similar entrance spaces, simply labeled "cable termination vault."

When properly outfitted, these installations have value in decoupling the exterior from the interior environment and in reducing the secondary effects of nonlinear operation of the protective devices themselves.

When an EMP entry vault is utilized, it should be large enough to house all the necessary terminal protective devices (filters and arresters). Further, the output leads from the terminal protective devices should be isolated (filtered) as they enter the shielded enclosure by means of feed through capacitors.

Circumvention Techniques

Previous paragraphs in this section have discussed a variety of approaches for reducing the amplitude or frequency spectrum of EMP induced waveforms. Through proper design, these approaches can reduce these induced transients to a level sufficiently low to prevent damage even to very sensitive components. However, in many cases, further reduction of the level of these transients to prevent circuit upset is neither cost effective or practical. Therefore, it is often necessary to find alternative approaches to circumvent the problem.

Circumvention approaches can be implemented either through hardware or software protection schemes. The hardware approaches can generally be categorized as either threat specific, or non-threat specific. The software approaches are generally non-threat specific. It must be remembered that these approaches are usually redundant to damage protection approaches since the susceptibility of most components is only slightly different if they are biased or not.

System Constraints

If we "gate down" a system in real time, there is at once an implication that the operational sequence execution would take place on a comparable timescale and with appropriate bandwidth. Only the more modern and sophisticated systems have such capabilities (i.e., 10 μsec stepping time). Such systems, however, are also open to a number of protection response options. For instance, the system response can be programmed to depend on where in the sequence the threat appears. It can overlook the threat if it is in a relatively invulnerable mode. It can stop and restart from some previously determined early stage or cancel a number of previous commands. It can similarly pause or hold, test for status validity, and start up again.

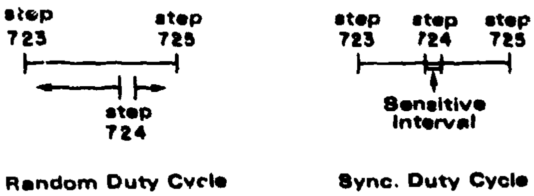
Older, or more primitive systems, generally cannot be desensitized "in time." Usually one must assume error or interruption, when a threshold field is reached, and simply restart the sequence. (this presumes that the system is hard enough to avoid permanent damage). In some cases, one may have the option of programming a separate sequence validity test which can negate or enable the mission sequence at some later stage.

Non-Threat-Specific Schemes

Duty-cycle schemes are generally permissible when the exact threat response time is not critical. If a particular system step requires 11 μsec to execute, but may be done anytime within 10 msec, then one may gain a reduction factor of 100 in threat coincidence probability by suitable cycle suppression.

Both random and synchronous schemes have been considered. The synchronous scheme lends itself to certain forms of bandwidth reduction as well. A variety of gating and switching techniques can be applied for disabling circuit inputs during the "off" periods. Redundant message transfer is another alternative.

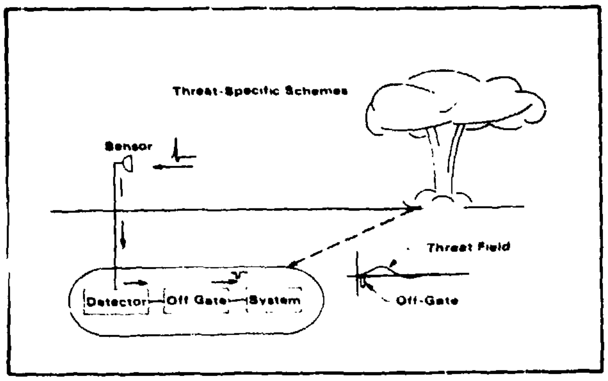
Non-Threat Specific Schemes



Threat-Specific Schemes

An active nuclear threat may be sensed in a number of ways. Let's confine attention to "prompt-spike detection." The basic reason that this works for EMP is the waveform peak inside a system is generally much broader and, hence, later than outside. In principle, one can use the exterior-sensed signal to "gate down" the execution sequence before the internal environment reaches error or interruption levels.

The biggest problem with this scheme lies in "false triggers." Experience indicates that it is almost essential to couple two different prompt sensors in coincidence in order to avoid almost continuous system inhibition due to non-nuclear noise. By "different," we really mean different in nature, such as an EMP antenna and a photoelectric unit.



Coding Techniques

Another approach to keeping unwanted energy out of sensitive circuits is to employ coding. A system which utilizes coded signals is far less likely to accept an EMP induced signal as correct information and respond to the interfering signal.

As stated previously, an EMP induced signal will normally appear as a damped ringing signal whose dominant frequency is system dependent. This signal, when processed by the system's, may appear as a series of positive, negative, or alternating pulses, etc.

The coding used for the desired signal must be sufficiently different from the EMP induced transient so it is not recognized by the system. Therefore, the coding must be of a different frequency, different polarity, different pulse groups, etc. Since this is highly system dependent, the system must be analyzed to determine whether coding is practical and, if it is, what code should be utilized.

Software Approaches

Circumvention can also be achieved in systems employing computer control through the system software. These programming approaches are usually based on error sensing and correction schemes, or through EMP event sensing as discussed previously.

The error sensing/correction schemes are usually based on a comparison of predicted data (allowable excursions/changes from one data sample to the next) versus the data collected from the systems sensors. If the collected data are outside of the prescribed bounds, the data are rejected. This type of scheme requires that the software program provides for a storing of at least the previous data sample. If the current sample is determined to be in error, it is rejected and the system "holds" on the previous data sample until the next data sample is received.

The use of threat sensing can work in much the same way. The system program in this scheme is also required to have a "store" and "hold" instruction in the program. If an EMP (or other transient) is detected, it is assumed that the data received during the time the transient occurs is in error and the system holds on the last correct data sample until new data is received.

Software programming of this type can be very effective in terms of hardening a system against circuit upset. It must be recognized that the computer memories (both volatile and nonvolatile) must be hardened against upset. This must be accomplished through the techniques already discussed elsewhere in this chapter. Among these techniques are shielding, terminal protection, and coding.

REFERENCES

"EMP Engineering and Design Principles," Bell Telephone Laboratories, Loop Transmission Division, Electrical Protection Department, Whippany, N.J., 1975.

"Electronic Pulse Handbook for Missles and Aircraft in Flight," Sandia Laboratories, AFWL-TR-73-68, EMP Interaction Note 1-1, September 1972.

"EMP Design Guidelines for Naval Ship Systems," IIT Research Institute, Chicago, Illinois. Naval Surface Weapons Center, Silver Spring, Md., NSWC/WOL/TR-75-193, August 1975.

Vance, E.F., "Design Guidelines for Treatment of Penetrations Entering Communication Facilities," Stanford Research Institute (for Defense Nuclear Agency), Menlo Park, Ca., August 1975.

SECTION VII

EMP SIMULATION, INSTRUMENTATION AND TESTING

7.1 INTRODUCTION

EMP hardness testing and protection verification necessitates use of experimental and analytical techniques for determining the response of systems, subsystems, circuits, and components to an electromagnetic wave.

Determining the response of a system, or a portion of a system, to an EMP is complicated by the geometry of the system and oftentimes poorly defined boundary conditions, that is, knowing the electrical properties of the various system components over the broad range of frequencies associated with the EMP. Thus, solving the problem of EMP effects on systems often requires experiments to determine system response. This experimental effort involves the use of transient sources, field illuminators, time domain instrumentation, and electromagnetic measurements and is the subject of this section.

Hardness testing or protection verification can be performed both in the laboratory and in the field. It can be performed on complete systems or portions of systems. EMP testing should be based upon sound physical laws, and the results should be rationalized in terms of those laws. Similarly, analysis should be compatible with the physical laws, and the results should be capable of experimental verification.

The analytical capability for EMP coupling is good provided the system can be adequately defined electromagnetically. Too often the simplifying assumptions which must be made to achieve an analytical solution do not account for subtle coupling modes which may be significant contributors to the system response. Testing can determine coupling and susceptibility for complex systems which cannot be rigorously analyzed.

Therefore, testing is essential for:

1. Verification of Analysis. The measurement of cable currents, voltages, and fields can verify analytical calculations of coupling modes, shielding effectiveness, and system response.

2. Extending Analysis. Testing can provide data on coupling, damage and upset thresholds, shielding effectiveness, coupling transfer functions, etc. The acquisition of experimental data can be a basis for improved analytical efforts.

3. Identify Weaknesses. Testing can quickly locate weak or susceptible points in the system that often can be hardened with simple, inexpensive modifications. The early location of critical weaknesses is very important for efficient EMP hardening design.

4. Verification of Protection. Testing can be used to verify that protective devices or techniques (filters, surge arresters, shielding, etc.) perform as required. This testing can often be performed at component and subsystem levels.

5. Certification and Quality Assurance. EMP hardness certification tests of the complete system increases the confidence that the system is hard. Component and subsystem tests can certify the design concepts. More often, tests will show margins or level of hardness. Repeat tests at all system levels can be used to assure the system or systems remain EMP hard even after modifications.

6. Life Cycle Hardness Assurance. Periodic testing can ensure that system hardness is not degraded due to environmental factors, modifications, retrofit, etc.

WHAT CAN BE THE OUTCOME OF TESTING?

- CONFIRM AND ASSIST ANALYSIS
- IDENTIFY WEAKNESSES
- VERIFY PROTECTIVE MEASURES
- CERTIFICATION AND QUALITY ASSURANCE
- LIFE CYCLE HARDNESS ASSURANCE

7.2 HARDNESS TESTING APPROACHES

The test philosophy and approach adopted are dependent on many factors, some of which are:

1. Threat Scenario. The threat scenario, i.e., the range of weapon type and yields are likely targeting information, the electromagnetic environment criteria, or the test criteria in the case of subsystems and equipments, provides the necessary information to determine the environment criteria to be simulated and assists in the selection of an appropriate simulator facility or approach.

2. Test Objectives. A complete definition of the purpose of the test, i.e., component damage factor, diagnostic information for design support, hardness evaluation, system/subsystem/equipment compliance to specifications or system certification, etc., is also required.

3. System Description. A complete description of the system is necessary to establish a meaningful test program. Information required includes a complete physical description (size and configuration), operational modes, system mission, detailed technical characteristics, subsystem/equipment criticality matrix to achieve the mission, etc.

4. Test Facilities. A knowledge of test facilities, both laboratory and field, is essential in establishing a test program. The availability, applicability, performance, etc., must be determined.

5. Logistics. In addition to the technical requirements, many logistics factors must be considered. These include costs associated with the test facility and fielding of the unit under test, scheduling of the test facility, test system availability, manpower requirements, and need for ancillary support such as special fuels, safety precautions, etc.

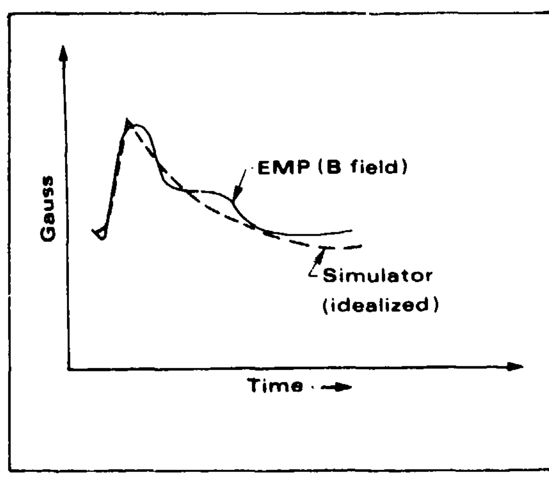
These factors are all part of the Hardness Test Plan. They must be spelled out in detail. Several hardness test plans are usually required to satisfy the overall Nuclear Validation Program even if only a single contractor is involved. Overall hardness assurance requires both analyses and test efforts throughout the design, development, test and evaluation phases of a program. Different approaches, techniques, procedures, facilities and instrumentation are required depending on the objectives of the specific test plan under consideration.

Simulation Requirements

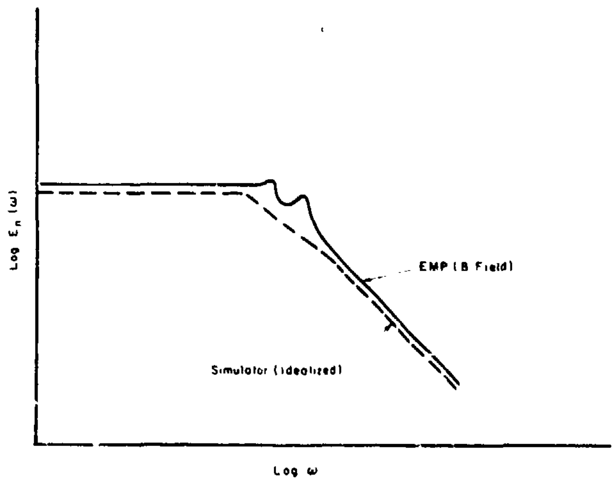
While it is desirable to test a system or portion of a system under actual conditions, an alternative approach is essential since this is not possible in the case of nuclear weapons effects. The alternatives are to simulate the EM fields produced by a nuclear detonation, or to simulate the coupled voltages and currents existing at the equipment terminals.

In order to use simulation for evaluation of a system or portion thereof, the first question that must be addressed is the required exactness of the simulation. Exact duplication of the environment, including any synergistic effects, is always a desirable goal. It must be recognized, however, that this is generally not achievable for either technical or economic reasons. Further, it must be recognized that exact duplication is not necessary to satisfy the goals of the test program.

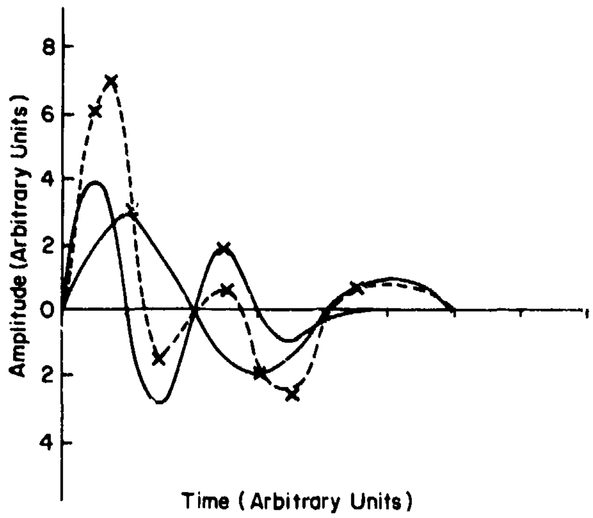
The important consideration, if testing is to be accomplished through the simulation of the EM fields, is that system response be determined over the entire frequency spectrum of the actual EMP or the specification criteria. Shown are typical waveforms depicting the specification criteria and an idealized output from an EMP simulator.



The specification criteria and the simulator output time histories do not exactly agree. However, if the frequency spectrum of the two are compared, it is apparent that they are essentially equivalent, that is, they have the same spectrum with minor differences appearing in the spectral energy density. These minor differences can be accounted for in the data reduction and analysis phase of the test program.



Similar considerations apply to the simulation of coupled voltages and currents at the equipment terminals. Most systems are of complex geometry. While in an idealistic view, the terminal voltages and currents will be a damped sinusoid at the resonant frequency of the coupling structure, this is rarely realized in the actual case. The systems, and consequently the coupling structures being complex, may contain several resonances resulting in a superposition of a number of damped sinusoid responses. Again, from a technical or economic viewpoint, duplication of the actual response may not be feasible. The alternative is to excite the system at several discrete frequencies and measure the response and combine the results through analysis.



Test Concepts

Simulation of an EMP criteria does not define a testing concept. Selection of a test concept must consider the factors stated previously. There are seven basic concepts which should be considered. These are:

1. Actual EMP Environment
2. Threat Criteria Simulation
3. Sub-criteria Coupling/Criteria Level Injection
4. Sub-criteria Coupling/Analysis/ Laboratory Injection
5. Low Level Coupling/Analysis/ Laboratory Injection
6. Analysis/Laboratory Injection
7. Scale Modeling

Actual EMP Environment

Exposure of the system to the actual EMP environment appears, at first glance, to be the ultimate choice. Testing of this type has been denied as a result of the Atmospheric Test Ban Treaty. Even if there was no test ban, if all threats as outlined in the threat scenario and all operating modes and configurations of the system were to be evaluated, the cost would be prohibitive. If only limited portions of the scenario were evaluated, analysis would be required to extend the test results to other situations. This form of testing would provide primarily a go-no-go test with only limited diagnostic information.

Threat Criteria Simulation

Testing the system at the threat criteria simulation, requires simulation of the actual EMP environment in terms of amplitude, time, and geometrical effects over the entire volume of the system to be tested. The system hardness evaluation is observed directly by the observed upset or damage or lack of system degradation. The margin of safety can be determined by exceeding the criteria level until damage or upset occurs. It should be noted that available simulators provide criteria level environments only over limited volumes.

This concept is simple and direct and may include such possible problems as nonlinear effects due to high amplitude fields, depending on the test configuration.

For comprehensive threat-level testing, the following requirements must be satisfied:

1. The orientation and direction of propagation for both the magnetic and electric fields must be simulated.

2. The time histories or spectra of both fields must be simulated.

3. The peak magnitudes of the fields must exceed criteria magnitude by the wanted margin of safety for EMP hardness, at least 20 dB and maybe more. Also, the relative magnitudes of magnetic and electric fields must simulate the criteria, i.e., the impedance of the field must be maintained. Specified criteria fields are not always radiated fields and, therefore, are not related by the free space impedance, such as source region fields.

4. The simulated electric and magnetic fields must be provided over a volume larger than the volume of the system to be tested.

5. The simulation must provide all possible criteria that may be imposed upon the system. If only one criteria is simulated, then the system is known to be hard to only that one condition. The relative location of the system with respect to the burst will determine the angle-of-arrival and polarization of the field. All angles-of-arrival and polarizations must be considered to ensure system hardness to all threats.

SIMULATION

* FIELD STRUCTURE
* TIME HISTORIES
* MAGNITUDES
* VOLUME
* ALL POSSIBLE THREATS
  * ANGLE-OF-ARRIVAL
  * POLARIZATION

For criteria-level testing, the operational modes of the system must be well understood to guarantee that the simulated field was imposed at the most susceptible time to cause upset or damage. Without some other additional information on system susceptibility, the simulated criteria will have to be imposed for each and every operational mode. The same is true for each and every possible system configuration.

System

* Operational modes
* Configurations
* Availability

The generation of high amplitude fields (10's of kV/m) requires the use of extremely high voltage (several megavolts) transient energy sources and efficient, large scale illuminators. This usually dictates that these tests be performed on a single shot basis with several minutes to several hours between shots depending on the complexity and reliability of the pulse source, system under test, and test instrumentation.

Sub-Criteria Level Coupling Testing

Sub-criteria level coupling testing requires simulating the criteria field waveform, angles-of-arrival and polarizations but at less than the criterial amplitude. It provides the coupling response of the system under test.

This concept is advantageous over criteria level testing in that it eliminates possible personnel hazards, reduces instrumentation problems and improves shot recycle time. Also, danger of system damage is reduced or eliminated. It simplifies illumination of larger working volumes since much lower energy densities are involved. The same generic types of transient energy sources (possibly at lower levels) and illuminators are often employed.

The system response is monitored at selected points in the system in terms of the voltage and/or current. Since lower than criteria levels of fields are utilized, these voltages and currents must be analytically scaled to criteria levels to determine system effects. This scaling usually assumes linearity within the system. This is the major disadvantage of this form of testing in that any nonlinear devices (such as spark gaps, etc.) are not exercised or evaluated.

Another disadvantage of this test concept is a loss of measurement sensitivity due to the lower level fields employed.

Sub-criteria level coupling tests can also be performed using repetitive pulse test sources. Repetitive-pulse testing involves exciting the system with a free-running train of pulses having the approximate shape of the environmental pulse, but of reduced amplitude and a repetition rate of 10 to 100 pulses per second. The excitation levels used are sufficiently low that personnel hazards due to the fields are minimized (i.e., sensor and recorder connections and adjustments can be made without shock hazard).

Repetitive-pulse testing has the advantages of flexibility and rapidity with which data can be acquired. Because the excitation is applied on a repetitive basis for extended time, the system may be probed to locate areas of unusually large response and to ferret out the sources of unexpected responses. Locating and evaluating such responses are the most important results of a susceptibility test program. Such probing, sometimes referred to as "point-of-entry" testing, can be conducted in one part of the system while formal preplanned measurements are being conducted in several other parts of the system, since operators of the various measurements systems may work independently at their own speed. In single-shot experiments, on the other hand, all measurement systems must operate at the speed of the slowest system, since all measurement systems must be readied before the shot is fired. Utilization of measurement equipment and operating personnel is, therefore, much more efficient because of this flexibility afforded by a repetitive pulse test.

In addition to the increased data rate and probing flexibility permitted by the repetitive pulse approach, the quality of the data is greatly improved. In the single-shot tests, the oscilloscope operator must guess at the sensitivity and sweep speeds (and ocassionally the beam intensity and graticule lighting) to use initially and make corrections on succeeding shots until the desired trace is obtained. In repetitive pulse testing, the operator can see the trace and make all these adjustments before he records the trace on film. Thus, the probability that a particular film exposure will provide a useable record of system response is much higher in repetitive pulse testing. Furthermore, the losses accompanying misfires, false triggers, and missed triggers are greatly reduced.

Another important advantage of repetitive pulse testing accrues as a result of the ease with which the measuring equipment may be debugged. Because the repetitive pulse response is available "continuously," it is easy to perform quality checks on the instrumentation -- such as disconnecting the sensor to determine whether the observed response is a system response or spurious pickup in the measuring equipment, reversing the sensor connection to determine that the response reverses, experimenting with equipment location and orientation, and investigating the influence of equipment ground and power connections. Such checks are extremely time consuming when conducted on a single-shot basis and are often overlooked until thorough analysis of the data suggests the necessity of investigating the quality of the measurements. With the repetitive pulse system, however, these checks can be made quite thoroughly and relatively quickly. In addition, because the same sensing and recording equipment is normally used for both the low-level repetitive pulse and the full-scale single-shot tests, the debugging done with the repetitive pulse need not be repeated for the single-shot testing.

Complementing these important technical advantages is a cost advantage over and above the operating economy related to the more efficient use of personnel and equipment. Because the repetitive pulse system operates on a lower voltage than the full-scale, single-shot system, the cost of the high-voltage power supplies, high-voltage switch, storage capacitors (or lines), and illuminator (e.g., parallel-plate transmission line or antenna) will be significantly lower. Where it is in some cases approaching the limit of the state-of-the-art to produce switches, storage units, and structures to operate at voltages in excess of several megavolts, such components are readily available for use at 100 kV or less.

ADVANTAGES OF REPETITIVE PULSE TESTING

* CONTINUOUS PROBING
* EASE OF ATTAINING DATA
* VALIDITY OF DATA
* COST

Low-level repetitive pulse testing has several limitations. The most important of these is the limited ability to detect very small system responses with the lower excitation levels. This limitation can be circumvented (with additional measuring equipment) by using signal-averaging techniques to measure very small responses. However, the use of such techniques eliminates many of the important advantages of the repetitive pulse method since in signal-averaging techniques, the oscilloscope operator cannot see the trace until the averaging has been done over many cycles. Thus, the operator must make "blind" settings much the same as in single-shot testing. Also, because considerable averaging time is required to obtain an acceptable trace, the data rate is much lower when signal averaging is required (although perhaps still higher than with single-shot testing).

Sub-criteria level testing is intended to use simulation of the threat EMP time history and, with simpler linear extrapolation, determine the effects at threat magnitude. The assumption of system linearity is not easily circumvented except by a criteria level test with measurements that are easily compared to extrapolated sub-criteria measurements. Indeed, because one cannot be certain that the system will not be upset by the threat-level environmental pulse unless the system is tested with such a pulse, the final test must always be a criteria-level test. Testing is generally more efficient, however, if criteria-level testing is limited to the final phase in which modifications necessary to prevent upsets are tested. The bulk of the testing required to locate upset signal entry paths and coupling mechanisms and to test the effectiveness of proposed modifications can be conducted more efficiently at low levels with repetitive-pulse techniques.

Simulation of the EMP threat time history, field orientation, angle-of-arrival, polarization and direction of propagation are as necessary for repetitive pulse testing as for threat-level testing. If the test fields differ in any manner from the threat fields, then additional extrapolation is required to determine the effects of threat fields.

Weaknesses

- Sensitivity limitations
- Extrapolation to threat criteria
- Some threat simulation still necessary

Low Level Coupling Testing

Alternatives to subcriteria level coupling testing are low level coupling testing techniques. Two techniques which may be used utilize field illuminators with pulse sources which do not simulate amplitude or the use of continuous wave (CW) sources.

The use of non-representative pulse sources requires that the spectrum of the pulse source contain all the significant frequencies associated with the actual EMP waveform. Such pulse sources must have fast rise times and waveshape characteristics to provide both the high and low frequency content of the desired spectrum. One generator of this type is a Delta Function generator. This pulse source has extremely fast rise time and very short (impulse) duration. The spectrum of such a pulse is essentially a uniform spectral energy density.

The major disadvantage of this approach is sensitivity. Since the spectrum is so widely spread, the spectral density is quite low even for very high voltage sources. This limits the use, assuming state-of-the-technology of measurement instrumentation, to systems with high coupling efficiency or low isolation. A second disadvantage is, like sub-criteria level testing, it requires linear amplitude extrapolation to criteria levels over the entire frequency domain.

CW Testing

Testing with CW involves exciting the system with a single frequency (a spectral line) and measuring the coupling response of the system (magnitude and phase) at this frequency. This can be repeated at a sufficient number of frequencies over a broad frequency range to define the system transfer function (ratio of system response to exciting signal). This transfer function can then be used with the environmental pulse spectrum or any arbitrary source pulse spectrum to compute the pulse response of the system.

There are several advantages inherent in CW system testing. Measurement of the system transfer function (magnitude and phase) gives a complete description of the system response, provided the system is linear and sufficient measurements are made to define adequately the transfer function throughout the frequency range of interest. Utilizing Fourier transform techniques, the frequency range of interest and the number of test frequencies can be determined for the pulse shape specified.

Transfer functions defined for horizontal and vertical polarization can be used to determine the threat response for any arbitrary polarization.

The line spectrum of a transmitted CW signal permits the use of a narrowband receiver or tuned voltmeter to measure system response. Interfering signals outside the receiver passband, whether generated or external to the system, are easily rejected. This feature is significant for measurements conducted in a high-noise environment, such as a weapon system with generators and power supplies running. For a given source power, the signal-to-noise ratio in a narrowband system can be much higher than in a broadband system with the same power. This results in a large dynamic range for CW testing with fairly moderate exciting power.

CW testing is extremely well suited to extended probing to locate an energy-coupling mechanism. A single-frequency magnitude comparison can be made quickly between circuitry points, or at a single point with external system cabling connected or disconnected. Where several frequencies are used, this CW technique is effective in determining system attenuation or the effects of filtering or hardening modifications over the frequency range sampled. This probing technique is similar to that discussed for repetitive pulse testing and is similarly well suited for field application.

Analysis of coupling in the frequency domain facilitates development of LPN network models to represent EMP source and black box transfer functions. Once the system has been reduced to lumped components, a multitude of time and frequency domain network, analysis programs, such as SCEPTRE, NET-2 and TRAFFIC, are available to perform extrapolation to the threat criteria.

Use of Frequency to Define Time Response

- **Defines response for all threats**
- **Sensitivity and dynamic range**
- **Rapid system evaluation**
- **Some analytical efforts are easier in the frequency domain**

There are several CW testing requirements that must be satisfied to conduct successful transfer function measurements. These are:

1. Definition of a transfer function requires phase measurements as well as amplitude measurements.

2. Time is required to obtain data at many frequencies. This time can reduce data acquisition to times similar to criteria-level testing.

3. The number of frequency samples required to produce transfer functions is a function of the bandwidth of the pulse spectrum which is related to pulse rise time and duration. The number of samples required may be as high as 500 to 1000 to define a transfer function that contains significant variation over three decades, and possibly 10 times as many frequencies may be needed if very high resonant behavior is expected, such as receiver responses. A typical strip chart recording of both amplitude and phase data is presented.

4. Data obtained in the field must be machine processed to interpret the effects at criteria level. Simple extrapolation does not give answers, except as noted, when probing at a fixed frequency. However, if the testing is being done in support of analysis of coupling, or verification of models (equivalent cir-

cuits), it is possible to record transfer functions in real time. It is still necessary to execute extensive calculations, which may be performed best by machine, to make any comparisons in the time domain.

5. CW testing, like repetitive-pulse testing, assumes that extrapolation to threat magnitude can be done linearly.

CW Testing

Requirements and Limitations

- **Amplitude and phase**
- **Number of frequencies**
- **Time**
- **Data processing**
- **Linearity assumption**

Laboratory Testing

Testing as discussed so far has implied testing the entire system or subsystem in an EMP-simulated electromagnetic environment. This method of overall testing of all portions of the hardness problem is very desirable. However, other forms of testing can be effectively used in an EMP hardness program. These are generally laboratory-type tests and consist of scale model tests, cable driving, determination of component damage and circuit upset thresholds, shielding tests, and stationary field tests.

Laboratory Tests

- **Scale model tests**
- **Cable driving**
- **Component damage**
- **Shielding**
- **Stationary field**

Scale Modeling

In many cases, it is not practical to test the entire system for either technical or economic reasons. One alternative is to scale model the system as has been done by antenna designers for years with good agreement (generally within ± 6 dB). Some of the reasons for resorting to scale models are:

1. Simulation test facilities are not available.
2. Equipments are not available, especially during design/breadboard phases.
3. The system to be tested is very large, a VLF station, for example, cannot be placed in available simulators or moved to them.
4. Dedication costs, such as testing a shipboard system *in situ* where the ship would have to be dedicated for the full duration of the tests, are high.

Scale Modeling

- **Facilities not available**
- **Equipment not available**
- **System very large**
- **System dedication cost high**

Some of the benefits which accrue from scale model testing in the laboratory are:

1. Sensor locations can be determined prior to full-scale testing.
2. Design modifications or cable routing could be evaluated prior to incorporation on full-scale systems.
3. Worst and best-case conditions for EM angle-of-arrival and polarization could be determined.
4. Analysis can be validated by performing the analysis using the scale model parameters.

It must be recognized that, because of the difficulty in introducing minute openings or poor bonds into models, and since these often control interior fields, the usefulness of modeling is ordinarily limited to the measurement of external fields, voltages, and currents.

ADVANTAGES:

- LOW COST
- CONFIRM DESIGN
- VALIDATE ANALYSIS
- DETERMINE WORST CASE THREAT
- AID IN SENSOR PLACEMENT
- EVALUATE DESIGN MODIFICATIONS

DISADVANTAGES:

- EXTERNAL FIELDS, CURRENTS, VOLTAGES ONLY
- ALL EFFECTS NOT INCLUDED

In general, the same test techniques as for full-scale tests apply to scaled models. The test sources and illuminators as well as the scale model of the systems must satisfy certain general relationships. These relationships are shown in the following figure. One important consideration in the scale model is that it is not always feasible to scale conductivity of the enclosure. In order to reduce losses, since the frequency is scaled up, one alternative is to scale the conductivity/wall-thickness product ($\sigma_s t_s = \sigma_a t_a$).

GENERAL RELATIONSHIPS

MODEL SIZE	$D_s \quad \frac{D_a}{M}$
FREQUENCY	$\omega_s \quad M\omega_a$
CONDUCTIVITY	$\sigma_s \quad M\sigma_a$
PERMITTIVITY	$\epsilon_s \quad \epsilon_a$
PERMEABILITY	$\mu_s = \mu_a$
WAVELENGTH	$\lambda_s = \frac{\lambda_a}{M}$
PROPAGATION LOSS	$\alpha_s = M\alpha_a$
PROPAGATION PHASE	$\beta_s = M\beta_a$

Direct Injection Testing

Current and/or voltage waveforms can be directly injected into the system cables or equipment terminals to determine upset/damage levels or the transfer function response of the system. The system configuration and test objectives will determine the required characteristics of the driving source.

Sub-criteria level and low level coupling tests can be combined with criteria level injection tests to evaluate system degradation effects. These tests involve injecting the waveforms monitored during low level field tests with amplitudes scaled to criteria levels. It is very difficult and costly to generate the complex waveforms often observed, especially at the criteria levels. An alternative, mentioned previously, is to generate a series of discrete frequency waveforms one at a time which cover the frequency spectra of the measured waveforms. To select the proper frequencies and waveform amplitudes usually requires careful analysis.

Transfer function measurements can be performed using either CW or pulse type sources. Pulse sources must have a frequency spectrum consistent with the anticipated or measured spectrum of the energy coupled to the system. CW sources, being discrete spectral line sources, must be available over the same spectrum. These pulse and CW sources are usually much less costly than coupling sources because this technique uses a small source tightly coupled to one port rather than a large source loosely coupled to the entire system.

Direct injection testing has the advantage that one or more ports can be studied singularly or in any combination. It is particularly useful when components or circuits must be subjected to transient voltages or currents to determine their damage thresholds and failure points.

The main disadvantage associated with direct injection testing is that the actual free field coupling to the total system cannot be simulated. Correctly phasing and shaping the pulses for a multiport injection system can be very difficult. Also, depending on the point of injection, any nonlinear effects may or may not be resolved.

Component Damage Testing

Components or circuits, both operational or in the breadboard stages, can be subjected to large voltages and currents to determine their damage or upset thresholds. The components which should be subjected to damage testing are those that normally would terminate long cable runs or antenna input cables.

Component damage tests provide a data base for use in predicting potential problem areas for both existing systems and those in the design stages. The data base is valuable as an aid in selecting components for use in new systems.

Based on component failure or circuit upset level, the amount of protection required to harden the system can also be specified for the postulated threat.

Component Damage and Circuit upset Tests

- **Establish data base**
- **Aid in component selection**
- **Establish protection requirements**

Shielding Tests

Shielding tests are most often performed at the subsystem level. Such testing can verify construction design and can locate energy penetration points before construction is complete.

Measurement of the shielding effectiveness of a facility or enclosure when constructed can also be used for hardness surveillance over the system life cycle. Periodic measurement of the shielding effectiveness will evaluate the condition of the shield by noting changes (deterioration) with time.

Analysis

Analysis must be used in combination with any of the test concepts. At criteria-level, testing analysis provides for extending the test results to other environments, system configurations, and system operating modes.

Pre-test analysis is useful for determining expected signal levels at critical EMP entry points.

When sub-criteria level testing is employed to determine coupling, analysis is required to scale the measured voltages and currents to criteria levels for direct injection testing where the correct waveform or simulation thereof is employed.

When low level coupling is employed, analysis provides for interpretation of the measured coupling transfer function in terms of the EMP environment criteria. Further, analysis must be employed to relate the laboratory tests on individual equipments to the environment criteria. These levels can then be compared and the potential for damage or upset assessed.

Analysis can also be employed to predict the coupling response of the system (see Section IV). The laboratory tests and additional analysis can then be employed as in the case of low level coupling testing.

7.3 TRANSIENT ENERGY SOURCES

Transient energy sources used in EMP hardness testing vary in complexity from simple commercial laboratory pulse generators used for component testing to sophisticated multimegavolt systems used in large scale simulators for complete systems tests. Standard commercially available pulse sources will be discussed briefly to aid the user in establishing a test setup.

An introduction to the various generic types of energy sources including the storage elements, switching arrangements, and the power supplies is provided. This discussion will be slanted toward considerations of the test facility user rather than the designer.

Energy Sources for Component and Circuit Testing

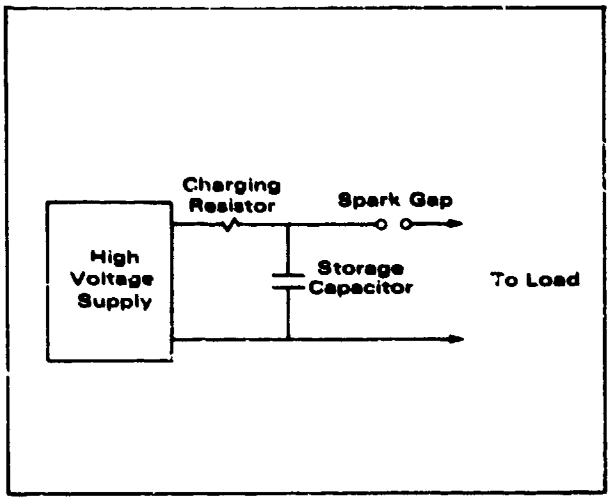
Component and circuit testing generally involves direct injection of current at the component or circuit terminals. The purpose of these tests is usually to determine upset or damage thresholds. As such, only low voltage/current pulse sources are required.

There are a wide variety of pulse sources available commercially with voltage ranges from a few volts to a few hundred volts (such as the Velonex Pulse Generator) and current ranges from milliamps to several amps. These units provide for variable output voltage, rise times, pulse widths, and fall times in many cases. These units are adequate for most component and circuit testing.

If greater power is required, these units may be used in conjunction with wide band (0-220 MHz) power amplifiers to faithfully amplify the pulse and preserve the rise time. These higher power units are usually available with selectable output impedances for matching purposes and current limiting provisions so the amplifier will not be damaged by failure tests where the failure may be manifest as a short circuit. Typical amplifiers which are available have 100 watt CW/400 watt pulse and 1 Kw CW/4Kw pulse capability.

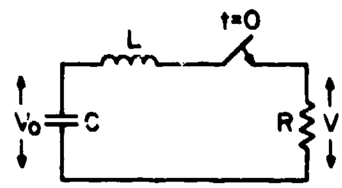
High-Level Energy Sources - Basic Energy Sources

The basic elements of an energy source for use with an EMP simulator or direct injection pulser are contained in a capacitive discharge pulser as illustrated.



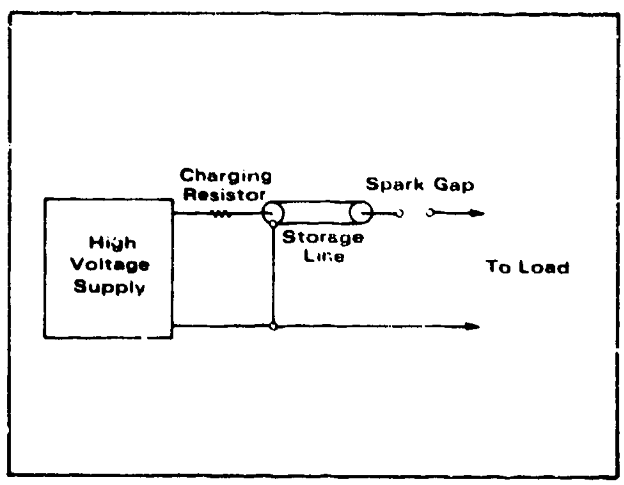
The source consists of a high voltage power supply, a storage element (capacitor), a current limiting charge resistor, and a switch for connecting the storage element to the load which can be a radiating structure, a bounded wave structure or the terminals of a subsystem under test. The decay time of the load voltage is determined by the storage capacitance value and the load impedance (resistance). The rise time of the applied voltage is controlled by the spark gap characteristics. Thus, the load, if resistive, will see a transient with a rapid exponential rise and a longer exponential decay. The charging rate is controlled by the storage capacitance, charging resistor, and current capability of the power supply. Normally, the storage capacitance is selected to give the desired transient decay time, and the charging resistor is selected for the desired pulse repetition rate.

The capacitor discharge pulser can be represented by the simple RLC circuit shown. The storage capacitor C is charged to an initial voltage V_0 and the switch is closed at $t = 0$ to discharge the capacitor into the load shown as a resistor R. The series inductance L may be the equivalent inductance of the capacitor and its connecting wiring or it may be a deliberately added inductance chosen to obtain a desired pulse rise time.

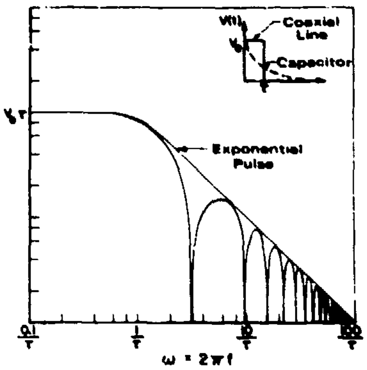


The load impedance is shown to be resistive which results in an exponential decay of a sinusoidal load voltage. The storage capacitance and load inductance determine the sinusoid frequency, and the resistance determines the envelope decay time constant or Q of the tuned circuit.

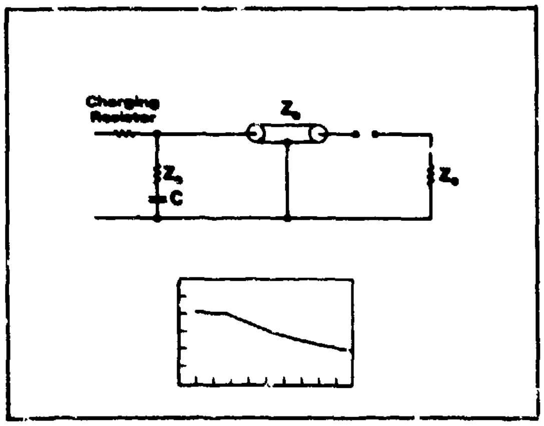
Another approach to obtain an energy source is to use a coaxial cable for the storage element. When the spark gap fires, the charged line voltage is divided between the line impedance and load impedance. The voltage wave travels down the line and is reflected by the charging resistor which is usually high compared to the line impedance. The reflected wave then returns down the line. If the load impedance equals the line impedance, then a voltage step of 1/2 the line charging voltage is applied to the load for a time equal to twice the line length. Note that the coaxial line storage element requires twice the charging voltage of an equivalent capacitor storage energy source.



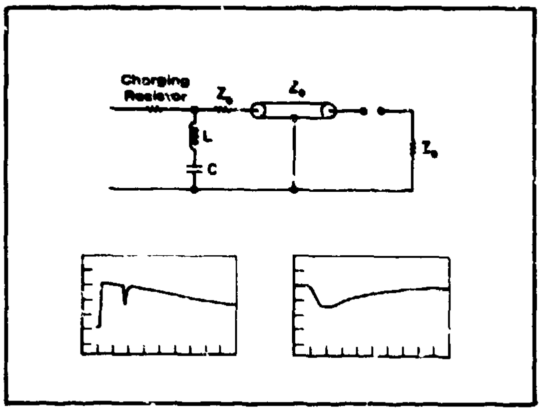
The pulse shapes and their spectral content differ for the two types of storage elements. The ideal charged transmission line matched to the load produces a rectangular pulse of width τ, while the lumped-capacitance source produced an exponentially decaying pulse of the time constant τ. The rectangular pulse from the transmission line has zeros at frequencies determined by the pulse width ($1/2\tau$), whereas the exponential pulse from the lumped-capacitance source contains no finite zeros in its spectrum. Because a real transmission line does not produce a perfectly rectangular pulse, the zeros of the ideal pulse spectrum will become minimal in a real pulse spectrum. Nevertheless, if a critical circuit or element is resonant at one of the null frequencies, it may not be excited sufficiently to produce the desired response one should acquire for the test.



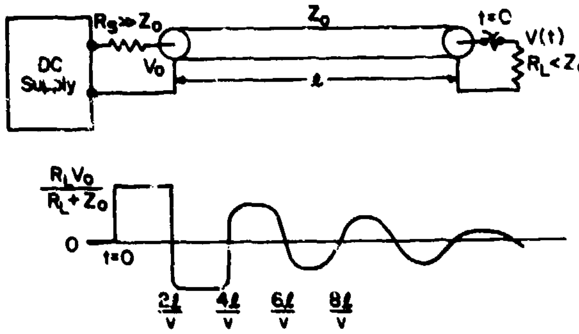
The square pulse that is inherent in the coaxial line storage system can be modified to improve the spectral content. With negligible inductance in the circuit, the storage capacitor energy dumped into the coaxial line produces a square pulse as is shown here. The resistance Z_0 in series with C prevents the capacitor from presenting the line with a high-frequency short that would cause an additional transient at four times the line length.



Noninductive capacitance at high voltages is difficult to obtain. The addition of an inductance to form an RLC network can be used to modify the coaxial line transient shape as shown here. The notch, expanded in the righthand waveform, is caused by the finite rise time restriction in the LC circuit. This notch can be smeared by using more than one coaxial line and using slightly different line lengths.



The charged transmission line can be used to produce a damped oscillatory waveform if the line is open circuited at one end ($R_s >> Z_0$) and its characteristic impedance is greater than the load impedance ($R_L < Z_0$). As illustrated, the waveform begins as a square wave oscillation and decays into approximately a sinusoidal wave after propagating a few round trips on the line.



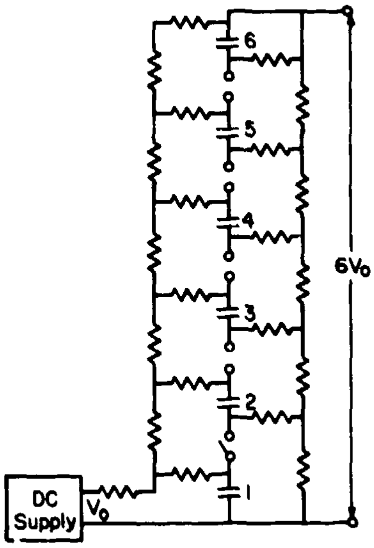
Charged Line For A Damped Oscillatory Waveform

Marx Generator

Generators are readily available which will produce voltages up to the ten's of kilovolts range. Larger voltages (megavolts) are obtainable by using special capacitor charging schemes. In a Marx generator, a bank of capacitors is charged in parallel. When the switches are closed, the capacitors are discharged in series. The series multiple gaps and their inductances limit the discharge rise time as discussed earlier.

Marx generator design has been improved so that these generators can be operated to produce waveforms with nanosecond rise times. The principle of operation of the Marx generator can be seen from the circuit diagram illustrated.

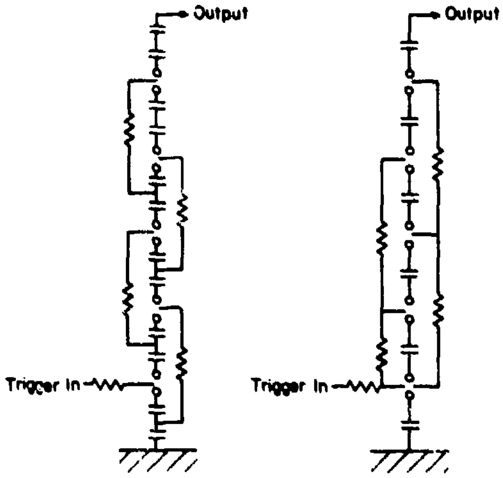
Each capacitor in the capacitor stack is charged to the dc power supply voltage (V_o) through the resistors alongside the capacitor stack. To erect the Marx generator, the switch at the bottom of the stack is closed to connect the bottom two capacitors in series, causing the voltage across the next switch to increase momentarily to well above its breakdown threshold. The second switch thus closes and overvoltages the third switch, and so on until all the switches are closed and the capacitors are connected in series. Thus, for N stages of capacitors charged to an initial voltage V_o, theoretically a peak voltage of NV_o can be obtained.



Basic Marx Generator Circuit Diagram

The time required to discharge such a generator is dependent on the stack length (number of stages), ionization times, gap inductance and load impedance of the gaps, etc. For megavolt range pulsers, rise times on the order of microseconds are typical.

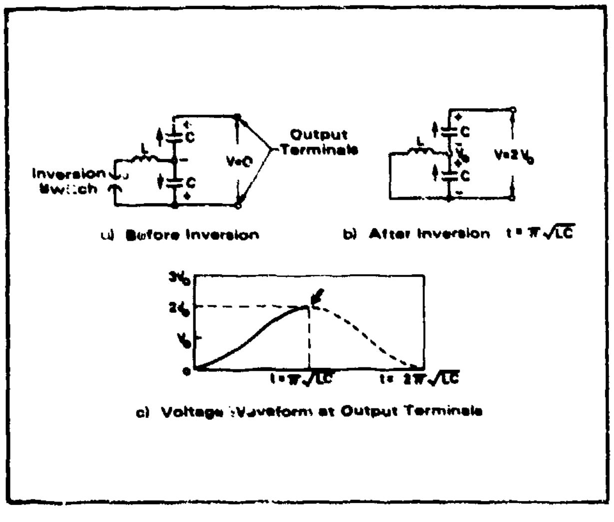
Fast erection of the stack is obtained by triggering the spark gap switches and reducing the size and inductance of the stack by developing high energy density capacitors and special packaging concepts. A typical Marx generator self-triggering erection scheme is illustrated below (the dc charging circuits have been left off to emphasize the transient circuit).



MARX GENERATOR SELF-TRIGGERING ERECTION SCHEMES

LC Inversion Generator

An alternate method of voltage enhancement is an LC inversion generator. Here, two charging capacitors are charged with opposing polarities. When the inversion switch is closed, the voltage on the lower capacitor will vary sinusoidally at a frequency determined by the LC circuit. When this voltage has reversed polarity, the voltage at the output is twice the charging voltage. Multiple sections of this generator can be placed in series to fire a single gap to the load. The ringing effect in the LC inversion generator is undesirable for some applications.



Van de Graaff Generator

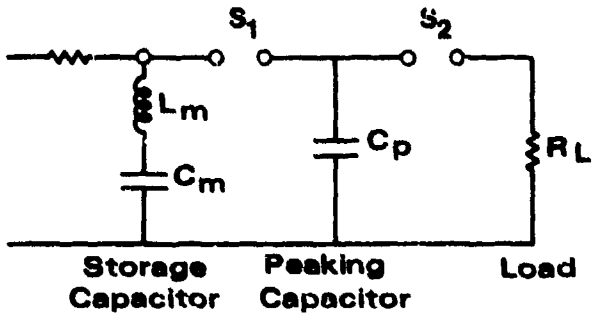
Another high voltage supply which has been used is the Van de Graaff generator. Supplies of this type are capable of generating extremely high voltages. Generally, since these generators utilize static charge transfer, the charging time is quite long. Supplies of this type find their main usage in criteria-level simulation of the single-shot type since charging times are too long to be utilized in a repetitive pulse type simulator where several pulses per second are desired.

Peaking and Transfer Capacitors

High-voltage generators that produce voltages in the megavolt range are physically large and usually contain too much inductance to deliver waveforms in the 1 to 10 ns range directly to the load. Consequently, some means of reducing the rise time of the pulse delivered to the load must be incorporated into the high voltage, fast rise time machines. The schemes commonly used for reducing the rise time of the waveform delivered to the load incorporate transfer or peaking capacitors at the output of the generators.

The peaking capacitor is a low value (compared to the erected stack capacitance), high-voltage, low-inductance capacitor that is capable of storing enough energy from the generator to allow the current to build up to the level required by the load. When the generator current has built up, the load is connected and the current is transferred to the load.

The peaking capacitor circuit is shown schematically, where C_m and L_m are the capacitance of the capacitor bank in the high voltage generator, C_p is the peaking capacitor and R_L is the load resistance.



If V_o is the peak voltage of the high voltage generator, the peak current in the load (for a zero rise time pulse) is V_o/R_L. The value of C_p is chosen so that when the voltage across it reaches V_o, after switch S_1 is closed, the current through it is V_o/R_L. At this time, switch S_2 is closed and the voltage across, and the current through the load immediately become V_o and V_o/R_L, respectively. Because the capacitance C_p, R_L, and the interconnecting conductors contain some inductance, the rise times of the voltage are greater than zero in practice; however, for the purpose of illustrating the principles employed, this stray inductance was neglected.

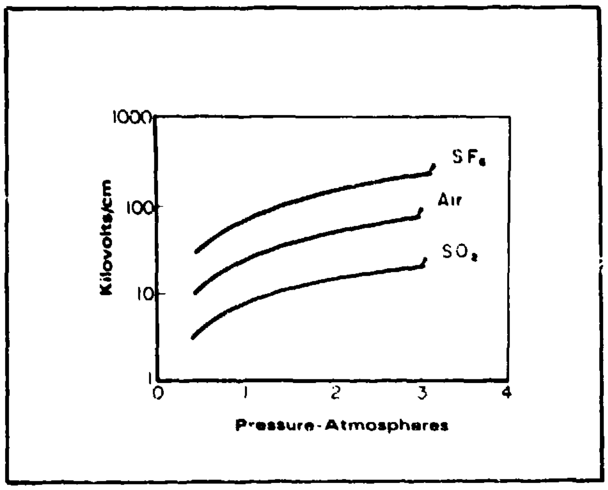
The advantage of the peaking capacitance method is that it permits a fairly small, high-quality (low inductance) capacitance to be used to form the leading edge of the pulse, with the primary energy storage remaining in the high-voltage generator.

The circuit for a transfer capacitor is identical to that for a peaking capacitor. The difference between the two is that only a portion of the source energy is temporarily stored in a peaking capacitor, while all of the source energy is temporarily stored in a transfer capacitor The transfer capacitor must be a low inductance device if fast rise times are to be achieved. Because the energy is only stored temporarily, however, leaky high-energy density dielectrics, such as water, can be used to make compact, low inductance transfer capacitors.

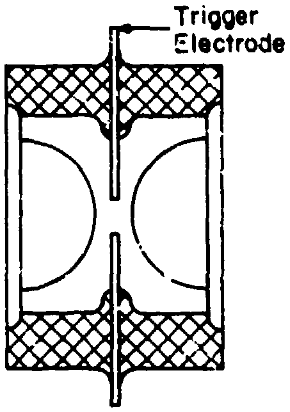
Transfer capacitors are used primarily in short-pulse systems in which the total stored energy is relatively small. Their advantage, compared to peaking capacitors, is that they can be used with load impedances that are neither constant, resistive, nor necessarily well defined.

Switches

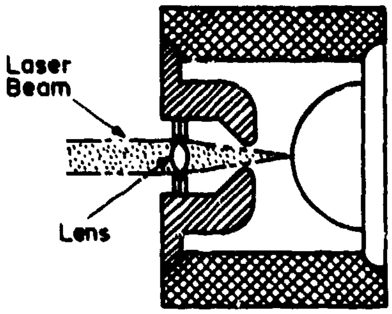
The switch inductance is the final limiting factor on rise time. The switch design is quite complex. A simple gap in air will fire when the voltage gradient reaches 20 kV/m at 1 atmosphere pressure. This arc will have a length and current that determine its inductance. The arc length (electrode spacing) can be reduced by pressurizing the switch and proper selection of the gas or gas mixture.



Spark-gap switches are often triggered to control the firing time. Two techniques are used. A laser can be used to produce ionization in the gap, or a pulse can be applied to a third electrode in the gap to initiate breakdown. This latter method adds additional circuitry into the active part of the system that must be included in the pulser design.



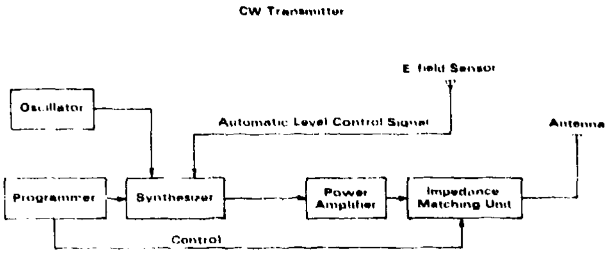
MID-PLANE TRIGGER



Axial Laser
Trigger

CW Energy Sources

The elements of a CW energy source are shown here. CW frequencies are selected from a stable source, amplified, in this case by a broadband amplifier, and matched to the radiating element. An alternate method is to use a tuned amplifier as the power amplifier. Generally, commercial equipment is available for CW energies.



7.4 ELECTROMAGNETIC PULSE FIELD SIMULATION

The EMP environment is usually defined as an electromagnetic plane wave propagating in free space. While it is not possible to exactly duplicate this environment, illuminating structures can be built which provide a good simulation of the environment over a large enough volume to obtain a valid assessment of a system.

There are two basic types of simulators commonly encountered in EMP testing. These two types are bounded wave or transmission line structures, and radiating antenna (monopole or dipole) structures. Other types of simulators are usually variations of these two basic types.

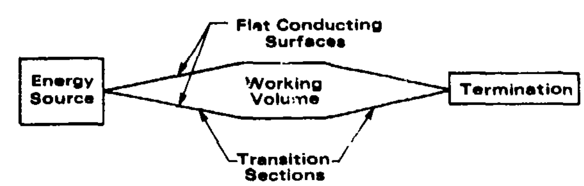
Bounded Wave Simulator

One efficient, broad bandwidth illuminator that is commonly used and is simple to construct is a transmission line simulator. This illuminator is a bounded-wave structure that is based upon a strip-line transmission line.

The principle of operation of a transmission line EMP simulator is to guide an electromagnetic wave across a test object situated between the two metallic surfaces of a transmission line. The essential elements of this simulator include an energy source, transition sections, a working volume, and a termination. An electromagnetic wave is generated by the pulse being applied to the transmission line and allowed to propagate to the terminal end. To provide the connection to the pulser and termination, transition sections of constant impedance are utilized. The cross-sectional dimensions of the working volume (i.e., the separation and plate width of the parallel metallic surfaces) must be large enough to provide a specified degree of field uniformity over the test object. A termination is provided to prevent the reflection of the guided wave back into the working volume. This termination is generally achieved by means of a transition section that guides the wave to a geometrically small resistive load whose impedance is equal to the characteristic impedance of the transmission line structure. Thus, for a transmission line simulator, most of the available energy is confined or bound to the space within the line so that large-magnitude fields are easily attainable.

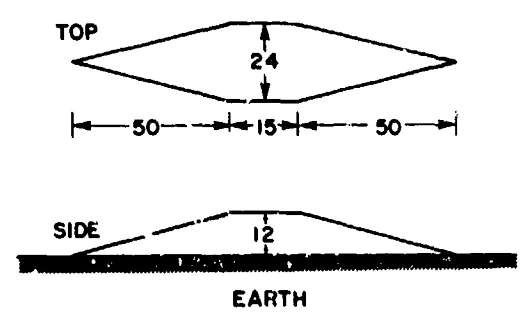
Bounded wave (transmission line) simulators are generally capable of generating EM fields at the EMP specification criteria level. The wave impedance in the simulator is that of free space (377 Ω). Their principle application is for missiles and aircraft in flight since ground effects are not present in these simulators. Also, they have limited interaction volumes, produce only a single polarization (normally vertical), and single angle of arrival. Different polarizations and angles of arrival are achieved by rotating the system under test which again places a restriction on system size that can be conveniently handled.

Components of a Transmission Line Simulator



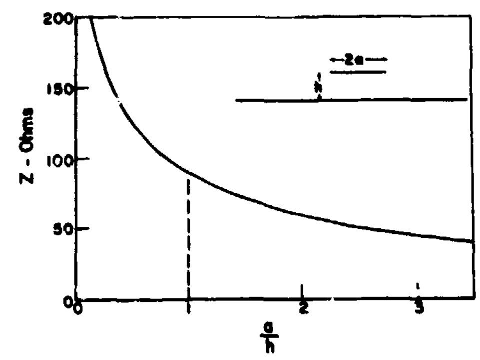
Let us consider some of the characteristics of transmission line simulators by an example. Shown is a transmission line with a 15 m x 24 m x 12 m high working volume. The line is over a flat ground plane and has transition sections that are 50 m long.

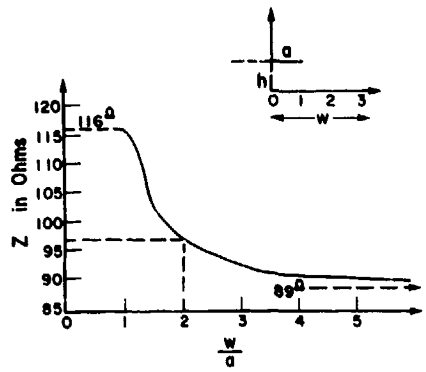
The impedance of the transmission line is a function of the width-to-height ratio which is constant from source to termination.



For an upper plate width of 2a at a height h above a perfect ground plane, the line impedance varies from 180 ohms for a/h = 0.2 to 40 ohms for a/h = 3.5. For our example, if the ground plane is a perfect conductor and the working volume has a ratio of a/h = 1, the impedance of the line would be 89 ohms.

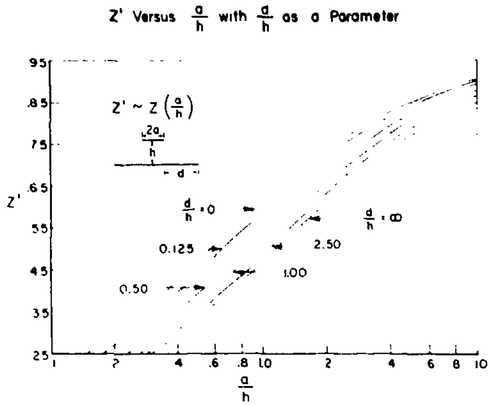
This impedance is for infinite width of the perfect ground plane.





As the width of the lower plate is decreased, the line impedance will increase. Shown is an impedance parameter, Z', that is proportional to line impedance, Z, and upper plate width-to-height ratio, a/h. As the width of the lower plate d is decreased for any a/h ratio, Z' shows an increase.

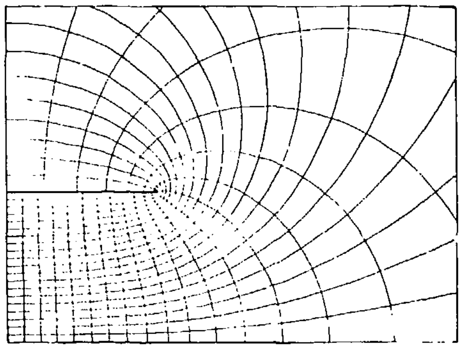
The electromagnetic field orientation and relative magnitudes for a strip-line transmission line can be displayed by a plot of equipotential lines and field lines. The equipotential lines show the magnetic field orientation, and their spacing relates to the electric field magnitude (volts per meter). The field lines on the plot show the electric field orientation, and their spacing relates to the magnetic field magnitude (amps per meter). The electric and magnetic fields are related by the wave impedance which is the free-space impedance of 120π or 377 ohms. For our example where a/h = 1 (Z = 178 ohm which is not the wave impedance), the field is essentially uniform under most of the volume covered by the upper plate.



For our example with a/h = 1, as the width of the lower plate (w) is reduced to the width of the upper plate (w/a = 1), the line impedance will increase from 89 ohms to 116 ohms. If we were to construct our example line with lower plate width twice the upper plate width (w/a = 2), then the line impedance would be 97 ohms. However, since the earth, which is a conductor, extends beyond the lower plate width, the actual impedance would be less than the indicated 97 ohms. For the presence of a real earth to have negligible effect, the lower plate should be four times the width of the upper plate.

Field and Potential Distribution for Parallel, Two-plate Transmission Line, 178.18 Ohms

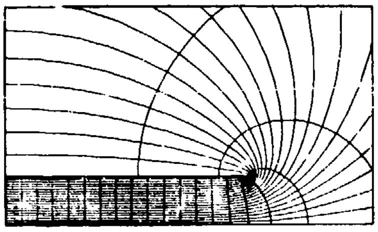
a/h = 1.00



The field uniformity is dependent upon the plate width-to-height ratio as indicated here for two extreme a/h ratios. For a/h ratios greater than one, the uniformity of the field is improved. For the 360-ohm line shown, a/h = 0.2, the electric field magnitude just below the upper plate is about 2.8 times the field directly below at the lower plate.

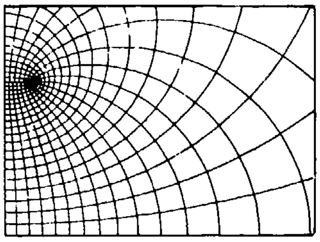
Field and Potential Distribution for Parallel, Two-plate Transmission Line, 57.97 Ohms

a/h = 5.00



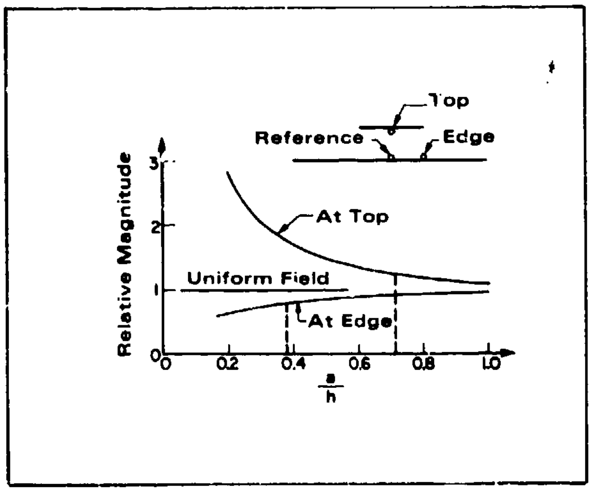
Field and Potential Distribution for Parallel, Two-plate Transmission Line, 360.08 Ohms

a/h = 0.20



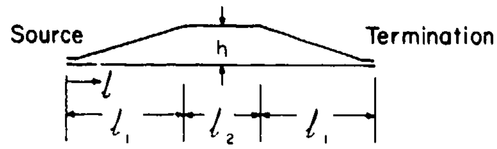
As a/h is reduced below unity, the fields at the top center and bottom edge of the working volume relative to the field at the bottom center vary as shown.

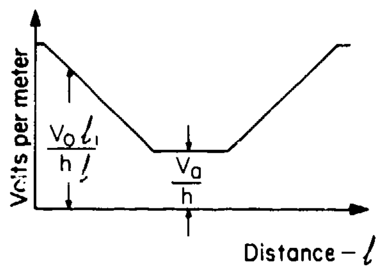
The field distortion is greater at the top center of the working volume than at the lower edge. This is indicated by the dashed lines which show the 20% variation points. For a maximum of 20% change in field magnitude at the top, a/h must be greater than 0.7, while the same distortion limit at the edge can be met for a/h = 0.4.



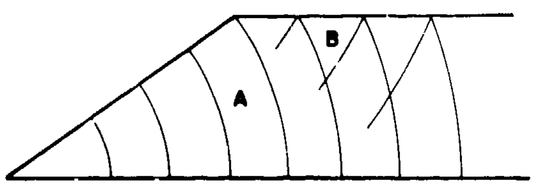
The fields wihin the working volume have been defined. These fields are essentially a TEM mode. For frequencies higher than the frequency at which the plate spacing, h, is half wavelength, it is possible to maintain higher order modes. However, since the wave is launched at a narrow plate spacing at the beginning of the transition section, only the lowest order transverse mode is launched. This mode travels down the transmission line and is not altered unless some discontinuity converts energy to the higher-order modes.

The magnitude of the field is a function of the applied voltage, V_o, and plate spacing, h. Thus, in the working volume the electric field is V_o/h v/m. The magnetic field is $V_o/h\eta_o$, where η_o is the free-space impedance (377 ohms). Within the source transition the electric field is $V_o\ell_1/h\ell$, where ℓ_1 is the length of the transition and ℓ is the distance from the source. The fields in the termination transition are similar. The maximum fields are generated at the source and termination.

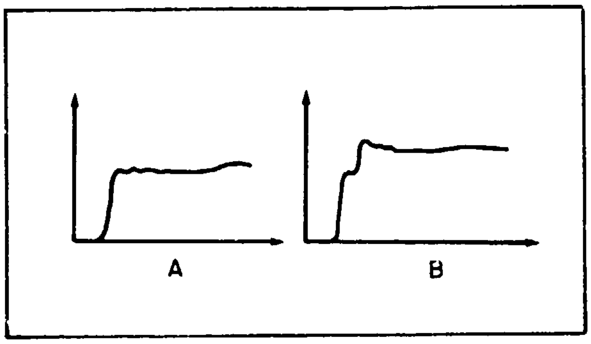




The wavefront in the transition section is spherical as shown. When this wave intercepts the working volume, reflections occur that will cause the field at B to be distorted. Fields at A should be a reproduction of the source signal.



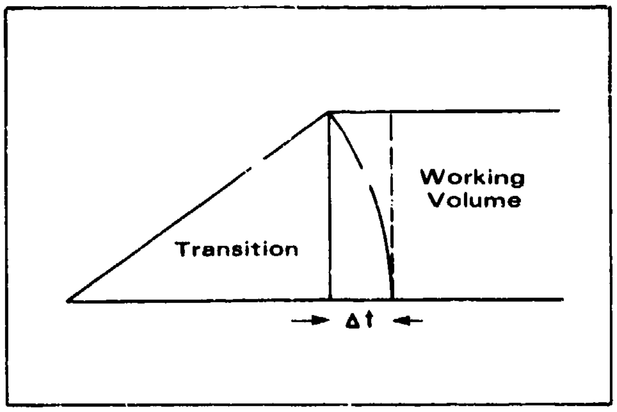
The accompanying figure shows two measured waveforms for a step input to the Alecs transmission line. A is the undistorted waveform. At B the reflected (and delayed) wave causes a distortion of the rise of the wave and alters the maximum amplitude.



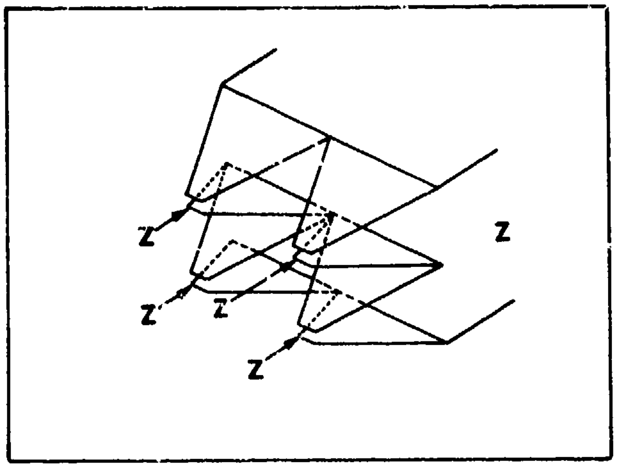
It is desirable that the field in the working volume be a plane wave as indicated by the dashed line. However, the wave from the transition section is spherical which means that the time of arrival of the wavefront at the top and bottom surfaces differs by Δt. The wave appears "tilted." For our example line with a height of 12 meters and a transition length of 50 meters, the time Δt is 5 nsec.

The wave tilt (Δt) is a measure of the time involved in going from the spherical wave to a plane wave in the working volume and is representative of the rise time of the line. A similar distortion occurs when the spherical wave intercepts the discontinuity at the beginning of the working volume at the edges of the upper plate.

The distortion at the mating of the transition section and working volume can be reduced by increasing the length of the transition section.

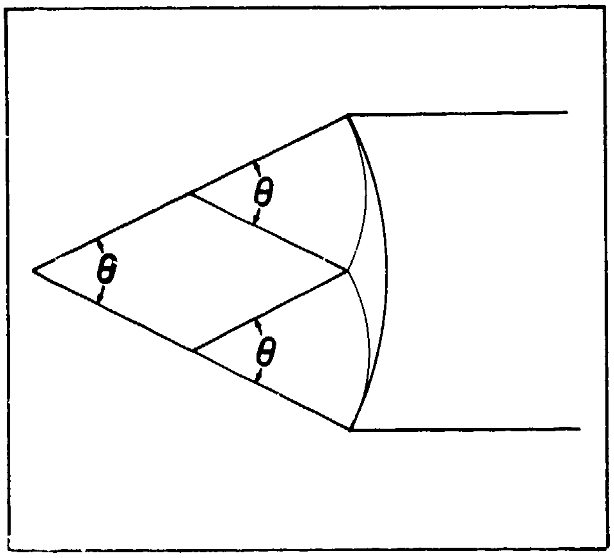


One method to achieve a more planar wavefront is the use of multiple transition sections as indicated here.



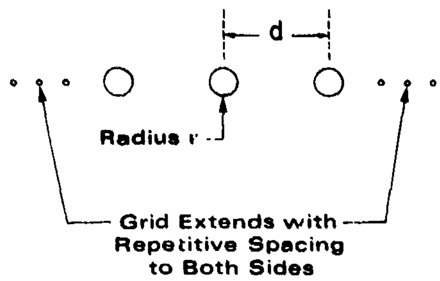
For the same length transition section, the spherical tilt is reduced. This improvement occurs even when the transition angle, θ, is held constant as indicated.

Multiple feeds force the spherical and plane wavefronts to be identical at the edges of the transition sections.



So far we have discussed a transmission line that is made of solid conducting plates. Portions of the line can be replaced with parallel wires if the spacing between wires is small compared to the highest-frequency wavelength. This use of spaced wires causes the field within a wire spacing to be distorted and will increase the line impedance. The effect of the parallel wires can be considered as an increase (Δh) in the vertical spacing between solid plates. For example, if the upper plate of our example line were made of #14 wire on 0.3 meter spacings (0.1 λ at 100 MHz), the increase in effective plate spacing would be 0.2 meter. For the 12-meter plate spacing in the example, the effect on the impedance would be small. For small transmission lines, the effect can be appreciable.

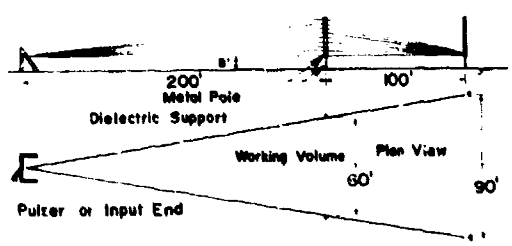
$$\Delta h = \frac{d}{2\pi} \ln \frac{d}{2\pi r}$$



For transmission lines made of parallel wires as shown, the plates are not perfect conducting plates, and at the higher frequencies will radiate energy. This radiation will cause a loss of high-frequency content in the wave and will maintain a spherical wavefront for high frequencies within the working volume. The transmission line can be crudely likened to a vertical rhombic antenna. The length of the line does not affect radiation from the line as long as the length is long compared to the height. However, as $\lambda/4$ approaches the plate spacing, the line will start to radiate as a vertical rhombic. For our example, a line with a plate spacing of 12 meters, frequencies above about 6 MHz will radiate, affecting high frequencies in the line.

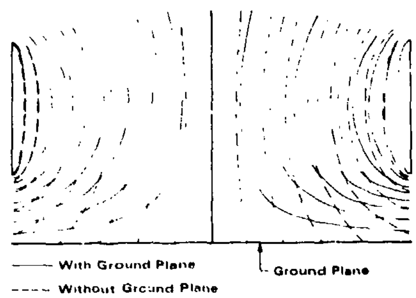
Transmission lines can also be used to create horizontal electric fields by erecting the plates vertically. Shown here is a concept that permits easy access to the work area. The work area is at the end of the source transition. The line termination is approximated by allowing the transmission line impedance to increase toward the free-space impedance of 377 Ω.

Horizontally-Polarized Simulator



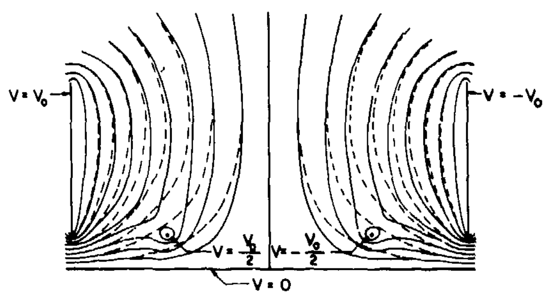
The field distribution within a horizontal transmission line is altered by the presence of the ground boundary on one side of the line. Shown are the equipotential lines with and without the presence of a perfect ground plane. The presence of the ground greatly alters the field orientation.

Field Distribution With and Without Ground Plane



This orientation can be improved by the use of wires to alter the field distribution. Shown is the effect of two wires at one-half plate potential. More complex configurations can make additional improvements.

A Field Distribution with Auxiliary Conductors

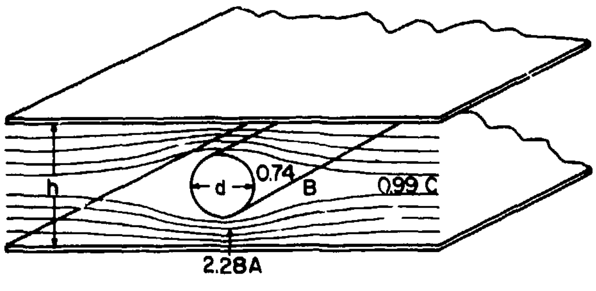


So far we have discussed the fields within a bounded structure (transmission line) without any conducting material (equipment) in the line. If the line dimensions are large compared to the equipment dimensions, then the field distortions caused by the equipment will have little effect on the fields created by the line. However, as is more often the case, the transmission line is made just big enough to contain the maximum dimension of the equipment being tested.

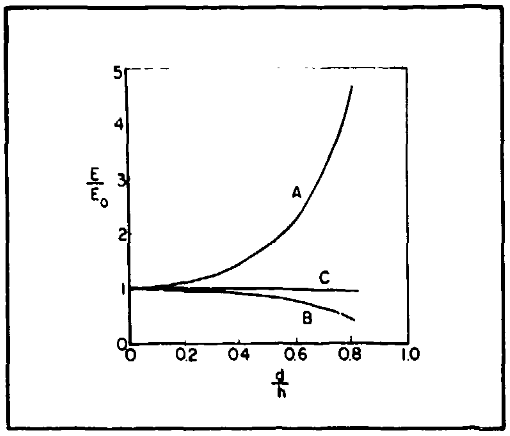
Placing a test object in the working volume can reflect and distort the incident wave. Large reflections directed back into the pulser might be objectionable if they either exceed the reverse voltage rating of the pulser or re-reflect from the pulser back into the working volume.

There will also be a certain amount of distortion in the vicinity of a metallic test object illuminated by a radiated plane wave. Providing the differences in field distortion between radiated plane wave illumination and a wave guided by a transmission line simulator whose cross-sectional dimensions are not significantly different from that of the test object are small, the use of the transmission line simulator is acceptable.

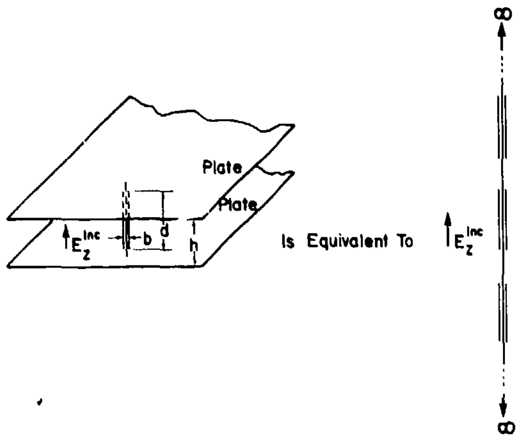
A conductor in a transmission line will alter the field structure in the line. Shown is a transmission line with an infinite cylinder that is 0.6h in diameter (d/h - 0.6). The equipotential lines can be seen to bunch between the cylinder and plates and are spread either side of the cylinder. The field at the upper plate (A in the figure) is enhanced to 2.28 times the undisturbed field in the line. At points B and C the field is 0.74 and 0.99 the undisturbed field, respectively.



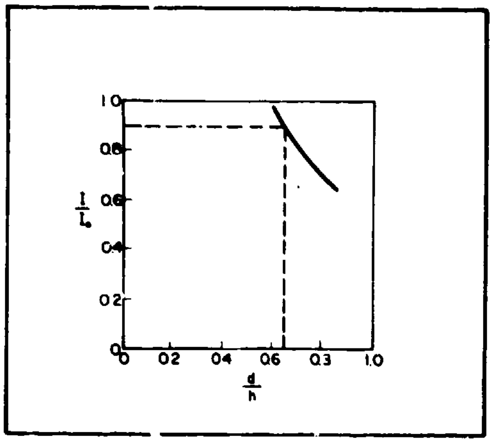
The field distrubance due to the cylinder for various d/h ratios is shown here. The maximum disturbance occurs at A. The fields at B near the cylinder are less altered. This enhancement of the field between the cylinder and plates effectively increases the transmission line capacitance and thus alters the line impedance. For a line with a d/h = 0.6 cylinder, the incremental line admittance is altered by 20%.

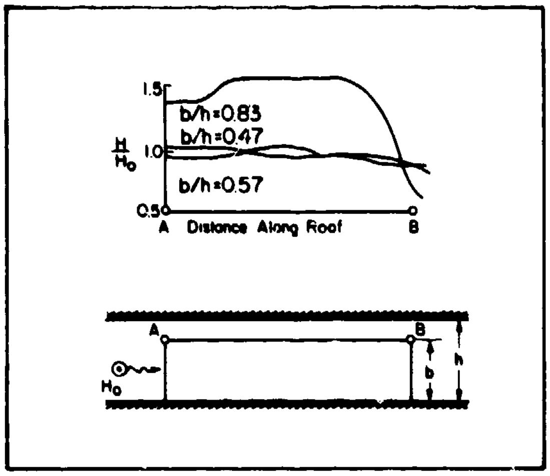


The effect upon the signal induced in a conductor can be approximated as shown. If a missile, approximated by a cylinder of diameter b and length d, is positioned between the plates of a transmission line as shown, an infinite series of images results. For d approaching h, the effect is to load the line, reducing the field in the vicinity of the cylinder.



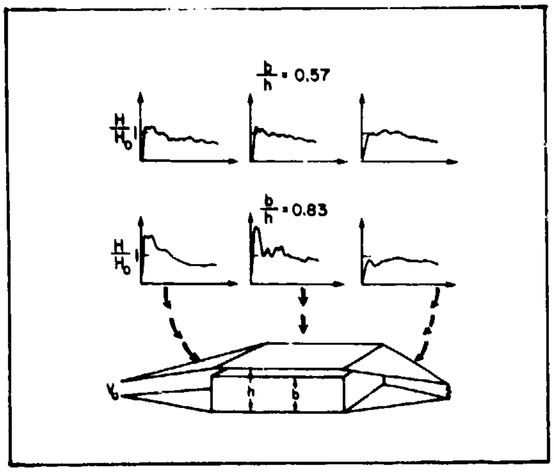
This loading reduces the peak current at the center of the cylinder as shown here. The reduction is less than 10% for d/h ratios less than 0.6.





The effect of a large object in a transmission line can be illustrated by the example shown here. Modeled is a transmission line constructed over a large rectangular building of height b. The transmission line width is the width of the building, and the building length fills the working volume. Shown are the relative magnetic fields for a step input to the line. For a building-to-line height ratio b/h - 0.83, the field at the front face and over the building is enhanced with high frequencies, while the high frequencies are attenuated at the back face. For a b/h ratio of 0.57 all fields are a fairly good reproduction of the incident step function.

These results can be explained using a transition section mismatched into a relatively low-impedance line formed by the upper plate and building roof. The resulting reflections give voltages on the line that result in the measured fields.



The peak magnetic field over the building indicates that, for b/h ratios less than 0.6, the transmission line can produce relatively uniform fields.

Pulsed Radiated Wave Simulator

An alternate method of subjecting a test object or system to a simulated EMP field is to use a radiating structure or antenna. A radiated-wave simulator radiates a free field which is not confined within the boundaries of the simulator structure.

An EMP-radiating simulator has several relative advantages. One is that construction is usually simpler and costs are usually less than those associated with a transmission line simulator. Another important advantage of a radiating simulator is that the space available to place test objects is not limited by the structure's dimensions. The disadvantage of such a simulator is that only a fraction of the stored energy is directed to the test object due to the relatively nondirectional radiation patterns of antennas used on existing simulators. Furthermore, there is a geometrical 1/R attenuation of the radiated wave amplitude with distance from the source. This geometrical attenuation causes a difficult tradeoff between high field intensities that can be achieved close to the antenna versus the nearly planar field distribution over large areas that can be obtained further away from the antenna.

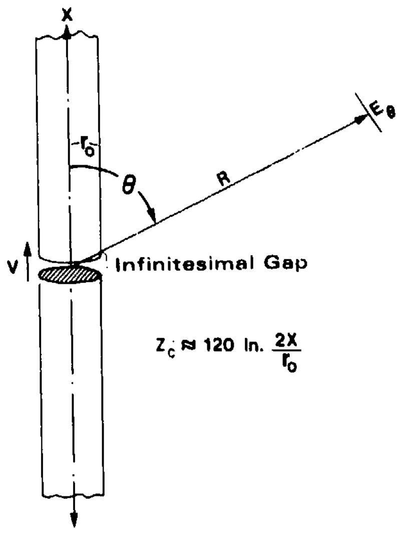
Pulse radiating simulators are generally of the biconic dipole or inverted conical monopole design. The dipole simulators are of the order of 1000 feet or more in length. Like the long-wire simulators, planarity over larger working volumes is obtained at the expense of field level (1/R falloff). The available polarization from the dipole facilities is predominantly horizontal on the line normal to the dipole axis and through the feed point of the biconic. Angle of arrival is also variable by varying the distance from the simulator. Ground effects must be considered. The clear time (time before arrival of the reflected pulse) is a function of distance from the antenna, decreasing as the distance increases. These facilities are applicable to ground based systems ranging from small vehicles to building size structures and ships. Aircraft can be tested in a fly-by mode.

The inverted conical monopole simulators are usually of the order of 100 feet high. Polarization is vertical and angle of arrival is fixed at grazing. As with all radiating simulators, planarity is obtained at a sacrifice of field level. Large working volumes can be obtained at reduced field levels. Ground effects are a loss of the high frequency content with distance from the simulator. These simulators are used where vertical field coupling is of interest.

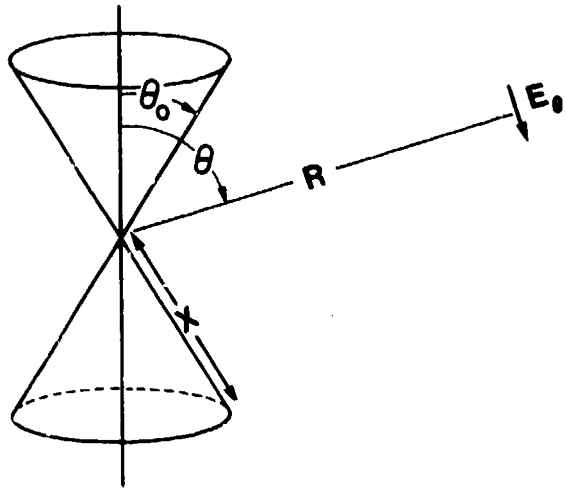
Biconic Antenna

The basic radiating antenna is a dipole. Shown is a cylindrical electric dipole of infinite length. The radiated field is zero in the axial direction (x) and a maximum normal to the gap. The magnitude of the radiated field varies as sin θ.

The radiated field time history depends on the current in the dipole elements. For a step voltage applied at the gap, the antenna current will be a constant voltage divided by the antenna impedance. This impedance increases logarithmically with the axial distance (x), causing the antenna current to decrease. The resulting radiated field for a step voltage applied to the antenna is a step followed by a logarithmic decay. For an applied voltage with a finite rise time, the increasing antenna impedance with x will degrade the radiated rise time.

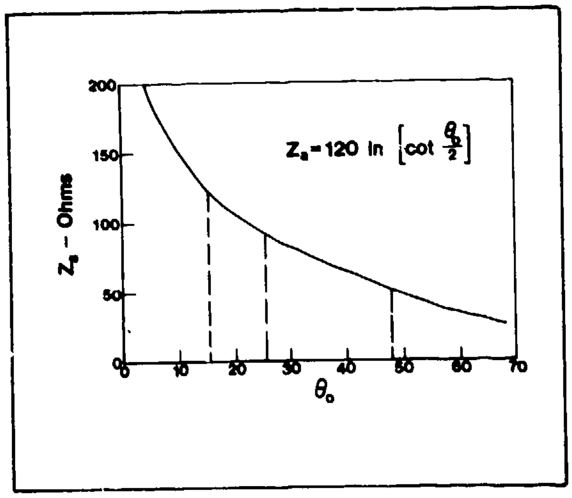


This feature can be overcome by the use of a biconic dipole shown here. The biconic dipole has two cone elements with a cone angle θ_o.



The biconic antenna impedance is a function of θ_o and does not vary with x. Typically, a 14° biconic has an impedance of 250 Ω, decreasing to 50 Ω at θ_o = 67°. This constant impedance with distance along the biconic results in an antenna current that is only dependent upon the applied voltage.

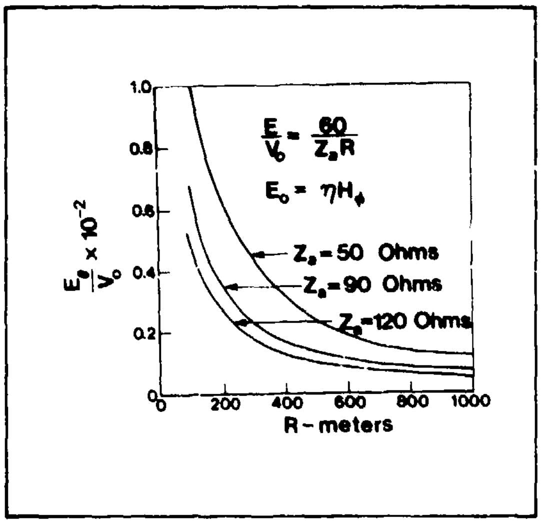
The fields emitted from a biconic (or dipole) have electric field components that vary as 1/R, $1/R^2$, and $1/R^3$. The 1/R fields are the radiated fields that are a direct function of antenna current.



The radiated electric field from a biconic is $E_\theta(t) = 60\, V(t - \frac{R}{c})/Z_a R \sin\theta$. For $\theta = 90^o$ (normal to the biconic), the radiated electric field varies as shown. The radiated magnetic field is related to the electric field (E_θ) by the free-space impedance ($E_\theta = \eta H_\phi$). This radiated field propagates from the biconic on a spherical wavefront. If the biconic antenna is long compared to the applied voltage rise time, then the rise time and peak magnitude are preserved in the radiated field.

The near-field components $1/R^2$ and $1/R^3$ must be considered near the biconic or dipole. The $1/R^2$ component varies as the time integral of antenna current and gradually is about 20% of the 1/R components. The $1/R^3$ component is even smaller. However, at distances close to the antenna, the magnitudes of the near-field components should be determined so that their contributions to the test object response can be assessed.

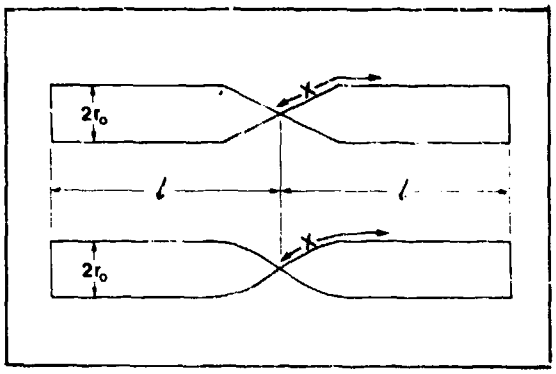
For applied voltages with long time durations, a biconic dipole soon becomes very large in diameter. Thus, biconic dipoles are normally used for only the rise time of a simulated EMP, thus limiting the length (x) of the biconic section.



For longer times, the biconic can be terminated in a cylindrical antenna as shown in the figure. The applied voltage rise time is radiated from the biconic section, and the later time signal is radiated from the cylindrical antenna. In this way, a pulse with a fast rise time and slow decay time (governed by the length of the cylindrical antenna) can be radiated. The applied voltage wavefront traveling along the axis will see the discontinuity in the radiated signal. This mismatch can be minimized by impedance matching the two antennas. Shown in the lower figure is an exponential match between the two antenna sections.

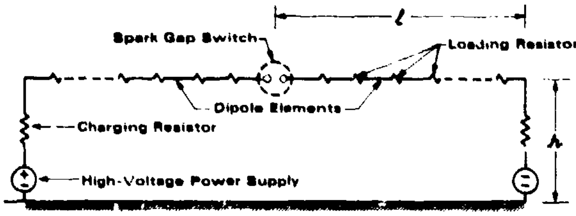
The length of the antenna must be long compared with the applied voltage duration so that reflections at the antenna end occur after the initial pulse duration. Thus, the radiated signal for a biconic-cylindrical antenna combination can duplicate the applied rise time and peak magnitude on the biconic section, followed by a radiated signal that depends upon the applied voltage and cylindrical antenna impedance. When this signal is reflected from the antenna ends, it will be reradiated from the cylindrical antenna. The reflected signal will follow the initial radiated pulse; i.e., if ℓ/c is larger than the applied signal duration, then the reflected signal will be time separable from the initial signal.

Reflections at the dipole ends can be eliminated by loading the cylindrical section of the antenna. This loading, which can be continuously distributed or distributed as lumped resistors, dissipates the applied signal before the antenna ends. An alternate technique often employed is to terminate the antenna with resistive elements to ground. This approach is valid since the propagation of the low frequencies on a dipole near the earth can be treated as a transmission line over earth and terminated with the effective transmission line characteristic impedance.



Resistive Loaded Horizontal Dipole Antenna

The biconic terminated dipole antenna described previously represents an exotic design that is capable of radiating a satisfactory simulation of a horizontally polarized EMP waveform. Such a design is required for many testing applications. However, for some applications a less complicated, easy-to-construct design may be adequate. The resistive loaded horizontal dipole shown is such a design.



Long wire dipole simulators are typically sub-criteria level simulators. They are usually on the order of 1000 feet long and, therefore, capable of illuminating larger structures. Planarity of the phase front is obtained at the expense of field level since the fields fall off as 1/R (R being the distance from the antenna). Angle of arrival can be varied also by moving out along the center line of the antenna. Polarization is predominantly horizontal on a line normal to the axis of the antenna passing through the center of the antenna. Ground effects and static fields must be considered in the use of these facilities.

To simulate an EMP, the radiating antenna must radiate at an adequately fast rise time and must be electrically long enough to radiate the entire waveform. One simple approach to eliminating the dipole end effects is by resistive loading of the cylindrical section of the antenna to distribute the end reflection along the entire length of the antenna. This loading, which can be continuously distributed or distributed as lumped resistors, dissipates the applied signal along the antenna and at the antenna ends.

The horizontally polarized resistive loaded dipole antenna is also called a "long wire" antenna and consists of resistively loaded horizontal dipole elements, two high-voltage dc power supplies, and a high-voltage spark-gap switch as illustrated. The two halves of the dipole are separated initially by the spark-gap switch and are slowly charged to opposite dc potentials (through a high charging resistance) from high-voltage dc power supplies until the spark-gap firing voltage is reached. The spark gap then fires, connecting the two oppositely charged halves of the dipole, producing a rapid, transient antenna current that is attenuated by the loading resistors as it propagates out from the spark gap. The resulting transient dipole antenna current is responsible for the radiated electromagnetic fields produced by the antenna.

The resistive loaded antenna is basically a symmetrical, horizontal dipole antenna drive by a step-function voltage source at the terminals. The resistive loading along the dipole elements attenuates the step-function as it is propagated along the elements, so that the reflection from the end of the element is small and the antenna does not ring. The radiated broadside field is thus a fast-rising pulse with a slow decay time, which is determined by the resistive loading. Because the resistive loading is lumped at intervals along the antenna elements, the rate of decay is not perfectly smooth. Furthermore, because the spark gap does not become conductive instantaneously, the applied voltage has a finite rise time that depends on the applied voltage rise time and the impedance characteristics of the first cylindrical section of the antenna.

The initial current I_o in the antenna can be determined from the applied voltage and cylindrical antenna impedance as in the previous discussion of a dipole antenna. This impedance is valid for times less than the clear time for the first ground reflection to be seen at the antenna elements. The initial current propagates along the first element with $I_o^2 Z_c$ = constant (Z_c is a function of distance) to the first lumped resistor, R_1, where the current is divided into a reflected current I_{R1} given by:

$$I_{R_1} = \frac{Z_1 - (Z_2 + R_1)}{Z_1 + (Z_2 + R_1)} I_i$$

and a transmitted current I_{T_1} given by:

$$I_{T_1} = \frac{2Z_1}{Z_1 + (Z_2 + R_1)} I_i$$

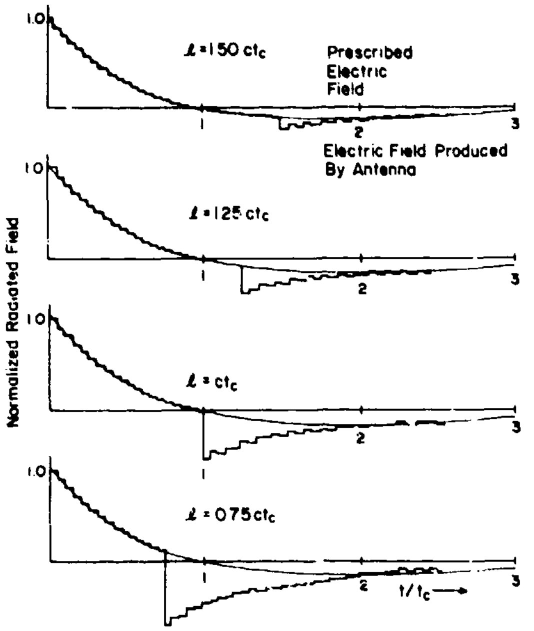
where I_i is the current in the first antenna element at R_1, Z_1 is the antenna impedance of the first antenna element impedance at R_1, and Z_2 is the second antenna element impedance at R_1. The transmitted current I_{T1} times the antenna impedance Z_2 creates a new voltage wavefront V_{T1} that propagates along the next antenna element. The reflected current causes a voltage wave to propagate back toward the gap.

This process is repeated at each lumped resistance as the wave propagates along the antenna elements. For later times during the waveform - after the time required for the first ground reflection from the gap to reach the particular lumped resistor - the impedances Z_1 and

Z_2 will be replaced by the single wire impedance over ground established by: $Z_T = 60 \ln 2h/r_o$ where Z_T = terminated transmission line impedance, h = height of line, and r_o = cylinder diameter.

The lumped resistors attenuate the antenna current gradually, so that current changes are relatively small and the current is practically eliminated by the time the current wave reaches the end of the long wire. Generally, 20 lumped resistors will provide a relatively smooth current waveform.

The resistive loaded dipole antenna radiates a field that is determined by the antenna current, which is controlled by the antenna resistors, by the antenna impedance and by the antenna length ℓ. The effect of antenna length on the radiated field normal to the antenna at the antenna gap for one synthesized antenna current is shown in the figure in normalized time. Two radiated field features are demonstrated in the figure. First, the reflected current at the ends of the antenna create a reflected field that increases as the antenna length is decreased, thus increasing the synthesized field overshoot. Second, the synthesized field crossover time must be less than ℓ/c if this overshoot is to be small. Generally, a ratio of antenna electrical length to crossover time of 1.4 will limit the overshoot to less than 20 percent.



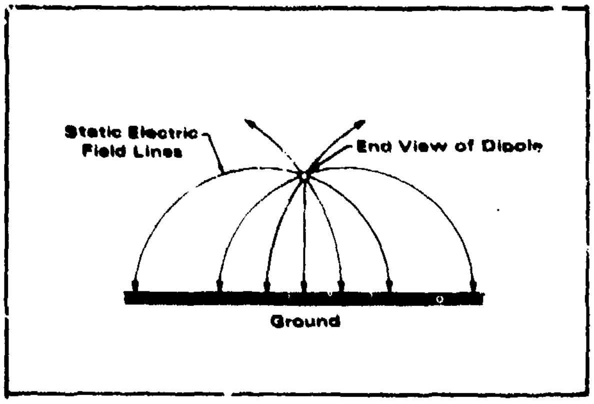
EFFECT OF FINITE ANTENNA LENGTH ON PULSE SYNTHESIS

The EM fields from a dipole antenna are radiated as a spherical wavefront with the origin at the antenna feed or gap. The electric field is in the plane containing the antenna and orthogonal to the line of sight to the gap. The magnetic field is orthogonal to the electric field and perpendicular to the plane containing the antenna. Both fields are at a maximum in the plane through the gap perpendicular to the antenna axis (θ = 90 degrees) and decline to zero in the direction of the antenna axis (θ = 0 degrees). The field magnitude is proportional to sin θ between these extremes and to 1/R from the gap (where R is the distance from the antenna along the center line).

For a horizontal dipole over the earth, the spherical wavefront will first intercept the earth directly beneath the gap. When the wave comes in contact with the earth, it will spread in a circular pattern along the earth. The magnitude of the E-field at the earth will be constant along a line parallel to the antenna axis. However, the arrival of the field along this line will be delayed away from the gap because of the spherical wavefront arrival along the line. The same phenomena applies along any line parallel to the antenna axis.

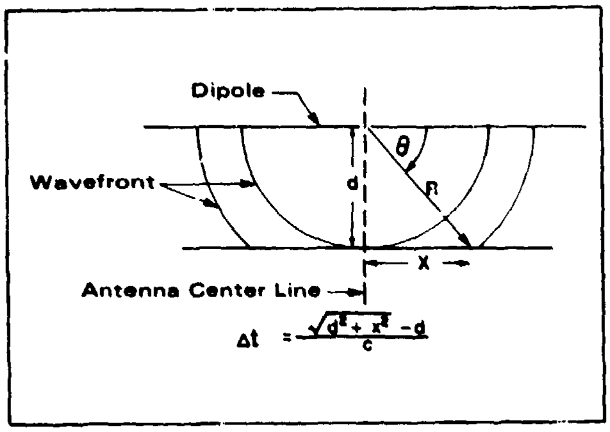
When the long-wire antenna is charged, a static electric field exists about the antenna which, at the surface of the ground, is vertically directed. When the spark gap fires, this vertical field begins to collapse as the discharge wavefront propagates down the charged antenna; a gradual change in the vertical field of the antenna results. The charge on the two elements is of opposite polarity, so that the static field of one element tends to cancel the field of the other. However, the cancellation is complete only at the center line, where both elements are equidistant from the observation point. Off the center line, the observation point is closer to one element than the other.

Therefore, the field of the near element is stronger than that of the far element, and the propagation time is shorter from the near element to the observation point than from the far element to the observation point. This difference in propagation produces a slight overshoot in the net field; otherwise the vertical field would have the appearance of a slowly rising step function. The variation of the vertical field strength with the position of the observation point is quite complex, however, because the field strength depends on the difference between two quantities whose magnitudes vary at the inverse cube of the range and whose times of arrival at the observation point also depend on the range.

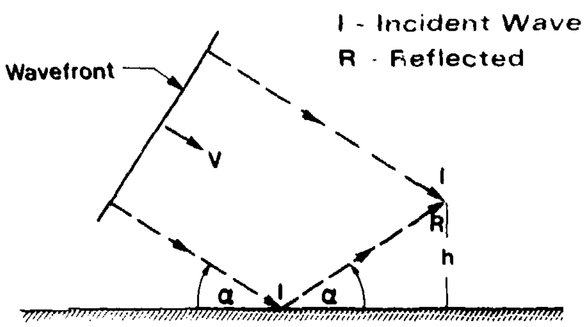


Ground Effects -
Horizontal Polarization

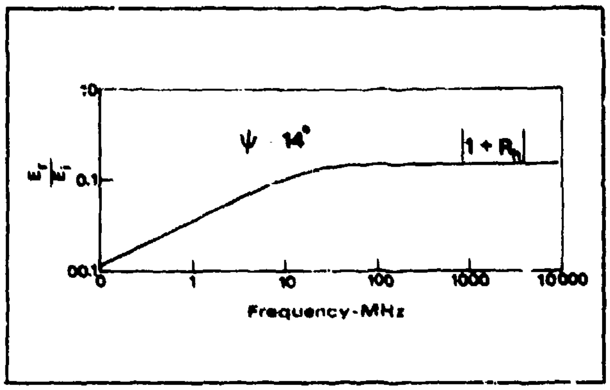
The long wire or biconic dipole antenna radiates a wave that expands as a spherical wavefront. This spherical wave differs from a plane wave in that, as the distance off the centerline (x) is increased, the wave arrives Δt seconds later. Thus, there is a time dependence and amplitude dependence for the fields impinging on a system near a dipole antenna.



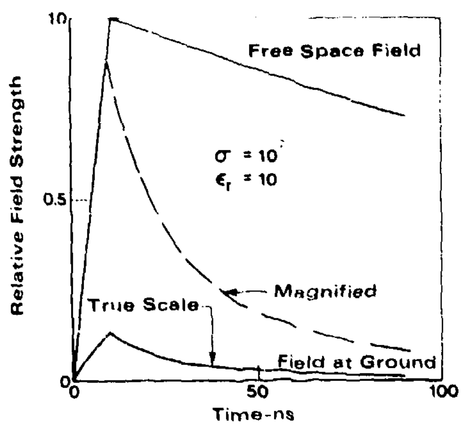
At the surface of the ground, the radiated fields are incident at an angle (α) that decreases with distance from the antenna. The presence of the ground reflects the fields which, in turn, recombine with the direct fields from the antenna.



The ground reflection (one plus the reflection coefficient) effect with frequency, shown here for a 14° incident angle, alters the field-time history. At short times (high frequencies), the earth appears to be a dielectric. At late times (low frequencies), the earth is a good conductor. The reflection coefficient (R_H) for horizontal polarization is negative (field reversal).

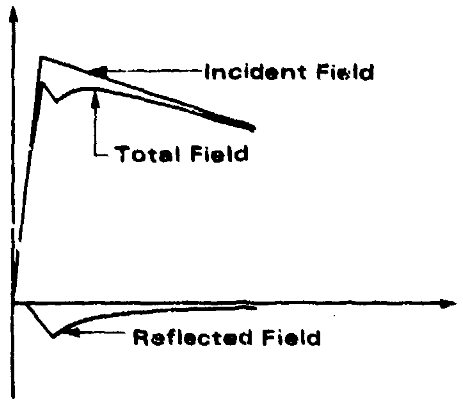


The effect on the electric field very near the ground is shown here. The high frequency rise time is preserved, and the low frequency content is decreased. The reflected wave is of lower magnitude than the incident wave and is reversed in polarity for the horizontally polarized component.



Thus, at a height above the ground, the electric fields add algebraically as shown here. The effect is a distortion of the pulse rise time.

Similar reflections occur for the magnetic field H so that the magnetic field above the surface is also distorted; however, in the case of the magnetic field, there is no phase reversal.



ELECTRIC FIELD ABOVE EARTH

Because both the horizontal electric field and the vertical magnetic field are operated on by the factor $1 + R_h$, neither of these components has the shape of the incident pulse. At the surface, however, the horizontal electric field induces a current in the soil (displacement current in early time and conduction current in late time) with an associated horizontal magnetic field. This horizontal magnetic field, H_r, is related to the total horizontal electric field, E, through the intrinsic impedance, η_s, of the soil.

For grazing incidence on a soil of high dielectric constant, this becomes

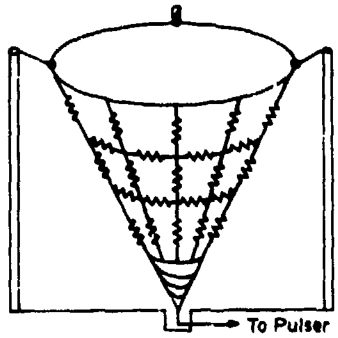
$$H_R \approx E_\theta \frac{2 \sin \psi}{\eta_o}$$

Since the dielectric constant of soil is generally large, the horizontal magnetic field at large distances from the antenna will have the same frequency dependence (and, therefore, the same time dependence) as the incident horizontal electric field. Hence, measurement of this component of the magnetic field will provide the incident pulse shape even through the principal components are distorted by the ground and by interference of the direct and reflected waves.

<u>Vertical Monocone</u>

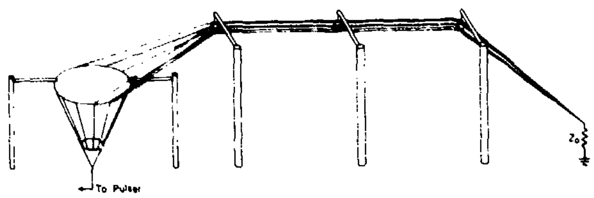
Another variation of a pulsed radiating illuminator to provide vertically polarized fields is the inverted vertical monocone antenna. A conic section antenna is utilized, as in the case of the biconic, to provide for the efficient radiation of the high frequency content of the pulse.

Due to construction difficulties and cost factors, the height of these types of antennas is usually limited to about 33 meters. This obviously limits the low frequency response to several hundred kilohertz (~ 900 kHz). To minimize reflections on the antenna, and consequently pulse distortion, the antenna is resistively loaded along its entire length, usually with discrete resistive elements, as shown.



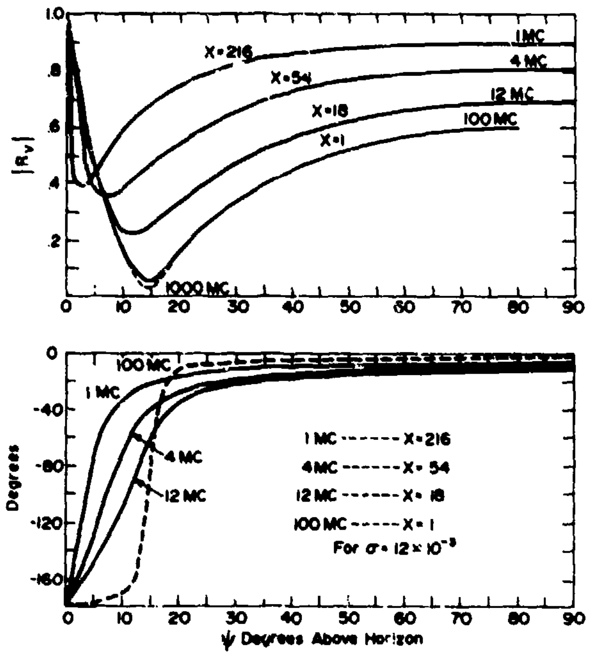
INVERTED VERTICAL MONOCONE WITH RESISTIVE LOADING

To improve the low frequency characteristics, longer tail on the pulse, one variation which has been employed is to use a terminated transmission line top loading on the antenna. The principal of operation is the low frequencies (<1 MHz) are provided by the fringing fields from the transmission line. The far end of the transmission line is terminated to earth with an impedance equal to the characteristic impedance of the transmission line to minimize reflections. This configuration is illustrated in the figure.



Ground Effects - Vertical Polarization

Radiation from a vertically polarized antenna (inverted monocone or monopole) is in the form of a ground wave. The reflection coefficient for vertically polarized waves at various angles of incidence (angle between the earth's surface and direction of the incident wave) is shown in the figure.



MAGNITUDE AND PHASE OF THE PLANE WAVE REFLECTION COEFFICIENT FOR VERTICAL POLARIZATION. THE CURVES ARE FOR A RELATIVELY GOOD EARTH ($\sigma = 12 \times 10^{-3}$, $\epsilon_r = 15$) BUT CAN BE USED TO GIVE APPROXIMATE RESULTS FOR OTHER EARTH CONDUCTIVITIES BY CALCULATING THE APPROPRIATE VALUE OF $X = 18 \times 10^{3} \sigma / f_{mc}$

The magnitude of the electric vector of the reflected wave at grazing incidence (ψ = o) is equal to the magnitude of the incident wave and has a 180 degree phase reversal. The magnitude of the reflection coefficient decreases with angle of incidence until the Brewster angle and then increases again toward unity. At grazing incidence the reflection coefficient is equal to one independent of frequency or ground plane conductivity. Above the Brewster angle the magnitude is a function of frequency and ground plane conductivity. It varies directly with a parameter x given by:

$$x = \frac{\sigma}{\omega \epsilon_v} = \frac{1.8 \times 10^{3} \sigma}{f_{mc}}$$

where

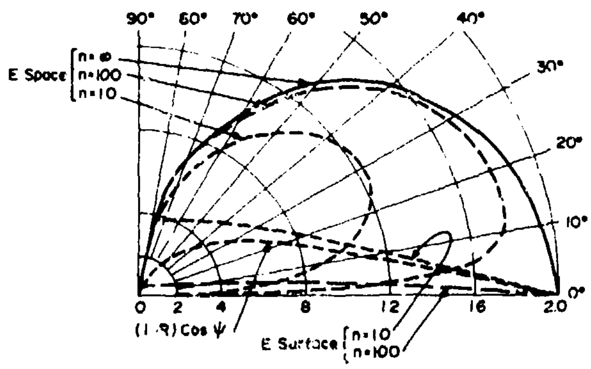
σ = ground plane conductivity
f_{MHz} = frequency in megahertz

The phase of the reflection coefficient varies from minus 180 degrees at grazing incidence to near zero above the Brewster angle being equal to minus 90 degrees at the Brewster angle. The result is destructive interference between the incident and reflected waves below the Brewster angle and construction interference above.

The ground wave can be divided into a space wave and a surface wave. The space wave consists of a direct and reflected component. When a radiating antenna is far above the ground plane (vertical dipole) the incident wave is plane and the ground wave and the space wave are identical. When the antenna (monocone) is on or near the surface, the incident wave is not plane and the reflected field must contain components in addition to those contained in the space wave. These additional components are the surface wave.

Because of the nature of the reflection coefficient, the direct and reflected fields of the space wave cancel for low angles of incidence for a finite conductivity ground plane. Therefore, the only field produced by an antenna on the surface of the earth (ground plane) is the surface wave which is not cancelled.

The radiation pattern for a vertical antenna on the earth's surface is shown in the following figure. Both the space wave and surface wave relative intensity is shown as a function of the incidence angle.



VERTICAL RADIATION PATTERN OF A VERTICAL DIPOLE AT THE SURFACE OF AN EARTH HAVING FINITE CONDUCTIVITY. THE PARAMETER $n = x/\epsilon_r$ AND AN AVERAGE VALUE $\epsilon_r = 15$ HAS BEEN USED BOTH SPACE WAVE AND UNATTENUATED SURFACE WAVE TERMS ARE SHOWN.

The surface wave has an attenuation factor associated with it that is directly proportional to the ground plane conductivity and inversely proportional to frequency and distance from the antenna. The attenuation is approximately equal to zero within a few wave lengths of the antenna and increases with distance. This factor is related to ground losses for ground planes with finite conductivity. The result is the high frequencies are attenuated whereas the low frequencies are not at the surface of the earth. Above the surface, the space wave dominates and this effect is not seen. Consequently, at or near the surface of the earth, the rise time deteriorates (is slowed).

These losses associated with the finite conductivity of the ground plane result in a horizontal electric field produced in the ground plane. This further distorts the wave in that the wavefront (E vector) becomes tilted.

These effects of a finite ground can be reduced by providing a high conductivity ground plane for the radiator. This could be of the form of a highly conducting metal mesh, such as copper clad steel mesh.

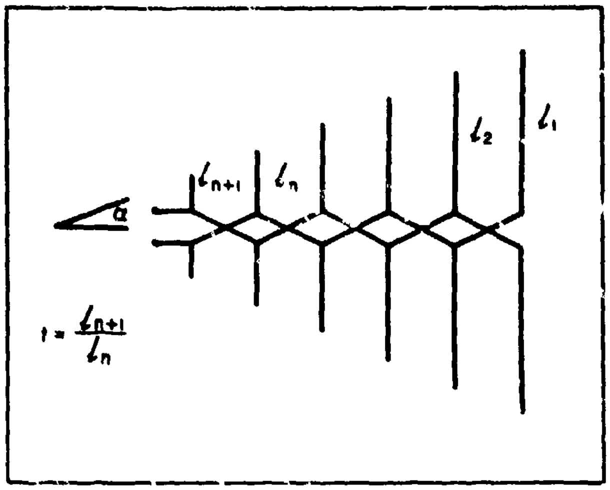
CW Radiators

The bounded-wave and radiating structures that have been discussed are broadband transient elements. For CW signals, some special forms of radiators can be used.

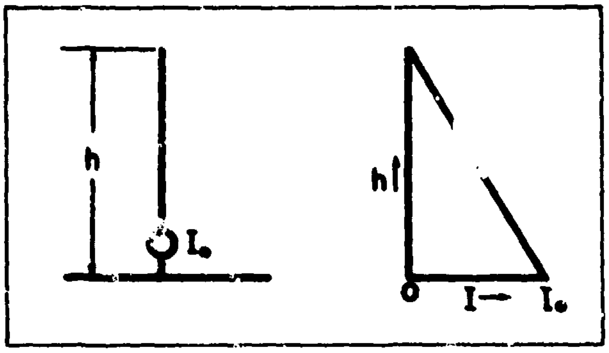
These radiators typically cover the frequency spectrum from a few MHz to 100-200 MHz due to limitations on the radiating structures. Antenna configurations are commonly log periodic types. Field strengths are usually to 1 v/m or less in most cases. Both vertical and horizontal polarization is possible.

These facilities can provide coupling response data as a function of frequency. To reconstruct the time response would require measuring both amplitude and phase.

The log periodic antenna is a CW antenna that can be used over a ten-to-one frequency range. The ratio of adjacent element lengths is given by t, and α is the array angle. These two parameters set the number or spacing between elements. Essentially, a log periodic antenna is a series of overlapping dipoles that do not vary their length rapidly. Thus, as the longest dipole decreases in radiation efficiency with increasing frequency, the next element increases. Thus, the phase center of the array moves toward the smaller elements as frequency is increased. Dipole elements are physically attainable for frequencies down to about 1 or 2 MHz. Log periodic antennas for horizontal or vertical polarizations are commercially available above 2 MHz.



For fields below 1 MHz, tuned electrically short antennas can be used. For example, a vertical monopole 100 feet long is resonant at about 2.4 MHz. Below 1 MHz, this antenna is capacitive and will have a current distribution that is approximately linear. The magnitude of the base current, I_o, can be maximized by tuning the antenna capacitance with inductance.



The fields of a vertical monopole are given by:

$$E_\theta = \frac{I_o h e}{2\pi} \left[\frac{j\omega\mu_o}{r} + \frac{1}{j\omega\epsilon_o r^3} + \frac{\eta_o}{r^2} \right] e^{-jkr} \sin\theta$$

$$H_\phi = \frac{I_o h e}{2\pi} \left[\frac{jk}{r} + \frac{1}{r^2} \right] e^{-jkr} \sin\theta$$

where:

I_o = Base current in amperes (with factor $e^{j\omega t}$ omitted)

h_e = Effective height of monopole in meters

ω = $2\pi f$, the signal radian frequency

μ_o = $4\pi \times 10^{-7}$, the permeability of free space

ϵ_o = $36\pi \times 10^{-9}$, the permittivity of free space

k = $\omega\sqrt{\mu_o \epsilon_o}$, the phase constant

η_o = $\sqrt{\mu_o / \epsilon_o}$, the intrinsic impedance of free space

j = $\sqrt{-1}$

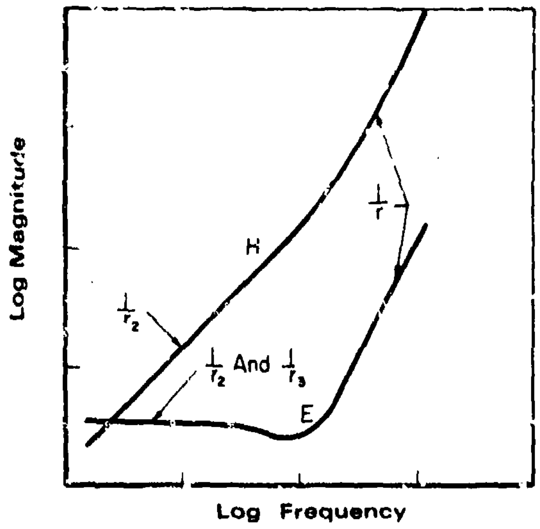
and r, θ, and ϕ are the spherical coordinates with the origin at the antenna base, and the axis of the antenna is the θ = 0 line.

The 1/R terms are the radiated far fields. Near the antenna, the so-called near fields, $1/R^2$ and $1/R^3$, must be considered.

The 1/R field term predominates after 3 to 5 wavelengths from the monopole. In the far field, E and H relate by the free-space impedance.

Near the monopole, a wavelength or so, E and H are no longer related by the free-space impedance. H decreases with frequency, while E is a relatively constant electrostatic field.

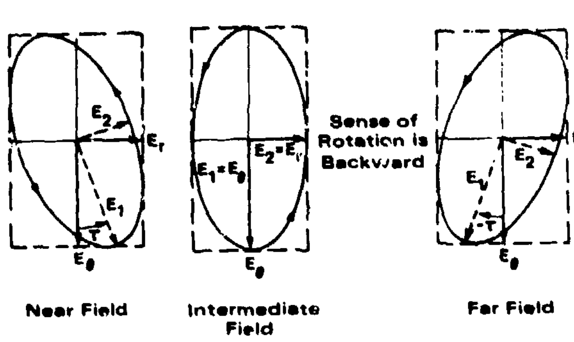
Field at a Fixed Distance



The fields that radiate from a monopole over real earth have an electric field in the direction of propagation (E_r). In the far field, the electric field is elliptically polarized with a tilt in the forward direction. Near the monopole, this ellipse is backward tilting. This tilting is caused by the change in time phase between the E-field components.

For low frequencies, the influence of the near-field components must also be considered.

Direction of Propagation



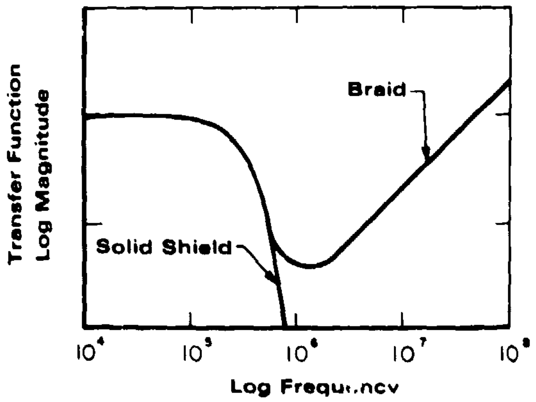
7.5 DIRECT INJECTION TECHNIQUES

Energy can be imposed on a system without creating electromagnetic fields in free space. Currents and voltages can be indirectly (cable drivers) or directly (direct connection, hard wired) injected into a system. The energy source to supply the current and/or voltage can be any of those described previously in the discussion of energy sources. The energy source and waveform chosen will be dependent on the objectives of the test program. This section will present alternative concepts for coupling the energy source to the subsystem, and/or equipment under test.

Cable Drivers

In determining an energy source for driving system or sub-system cables, consideration must be given to the minimal requirements in terms of both amplitude and spectral response. The amplitude of the shield or wire currents should at least be equal to the amplitudes derived from the coupling analysis or coupling measurements scaled to the specified environment criteria. This is true for either shielded cables or unshielded cables.

The spectral content of the energy source is another matter. For unshielded cables, the spectrum should agree with the coupling analysis or measured data. For shielded cables, however, the transfer impedance of the cable will modify the spectrum of the signal seen on the interior wires. The typical transfer impedances for solid shields and braided shields is as shown in the figure.

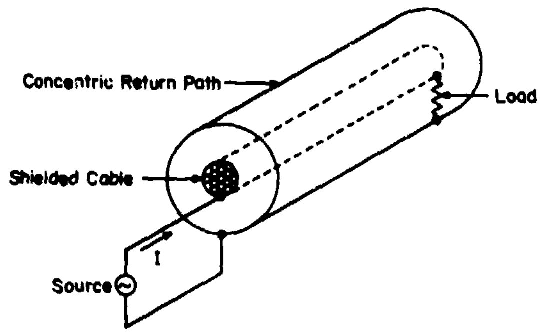


From the figure it can be seen that solid shields act like a low-pass filter. Braid shields, on the other hand, act like a low-pass filter until the frequencies are high enough to penetrate through the holes in the braid. Based on this typical response, a low frequency driver may be sufficient for solid shield cables but inadequate for braided shield cables. This must be verified, however, since this response must hold true for the cable including the cable connectors.

Direct injection techniques may be employed in either the time of frequency domain as stated earlier for field illumination techniques. To reiterate, if frequency domain measurements are employed, it is essential to measure both the amplitude and phase response if reconstruction of the time history of the pulse is desired.

Shielded Cable Driving Techniques

Cable shields can be driven by forming a transmission line between the cable shield and added conductors around the shield. This formed transmission line can be driven at one end with any pulse shape and terminated in its characteristic impedance. One of the simplest concepts for obtaining a uniform current density and characteristic impedance in a cable shield is to make the cable shield the center conductor of a coaxial transmission line as illustrated.



CONCENTRIC CYLINDER CURRENT INJECTION COUPLING STRUCTURE IN A COAXIAL TRANSMISSION LINE

With this configuration, the characteristic impedance (Z_o) of the transmission line formed by the cable shield, and its concentric current return path is:

$$Z_o = \frac{60}{\sqrt{\epsilon_r}} \log \frac{r_o}{r_s}$$

where

r_o = radius of the return path (cylinder)

r_s = outside radius of the shield

ϵ_r = dielectric constant of the material between the shield and return path.

As mentioned, the driving energy source can be an exponential pulse (capacitor discharge), a rectangular pulse, a damped sinusoid, etc. The exponential current pulse obtained from the capacitor discharge into a terminated transmission line has the desirable characteristics that (1) it is a fair simulation of the pulse shape that is induced in buried cables by the EMP, and (2) its spectrum is continuous - that is, it contains no dominant zeros or poles where the current spectrum is very small or very large. (For example, the rectangular pulse spectrum contains many zeros, and the lightly damped sinusoid spectrum is dominated by a narrow band). Although the flat bandwidth of the exponential pulse is $1/2\pi CZ_o$, (C is the capacitance of the discharge type source) its usable bandwidth is much greater because the spectrum is well behaved even when its magnitude is decreasing as 1/f. The usable spectrum is determined by the rise time of the pulse.

The capacitance, voltage, and characteristic impedance of a capacitive discharge source can be manipulated to obtain the desired spectral magnitude, peak current, or bandwidth within the ranges permitted by available capacitors and voltage breakdown limits of the coaxial configuration. The relationships for determining these characteristics of the source are:

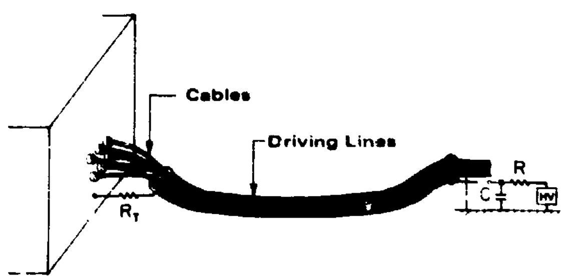
Bandwidth = $1/2\pi CZ_o$, in Hz

Spectral magnitude = CV, in ampere-seconds

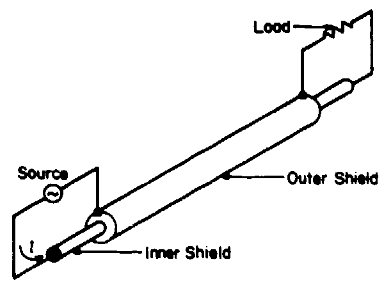
Peak current = V/Z_o, in amperes.

An approximation to the above concept of using a solid cylinder can be obtained by using a cage of wires. The number of wires required and their spacing is dependent on the highest frequency of interest in the source spectrum.

Cable Driver for Shielded Cables

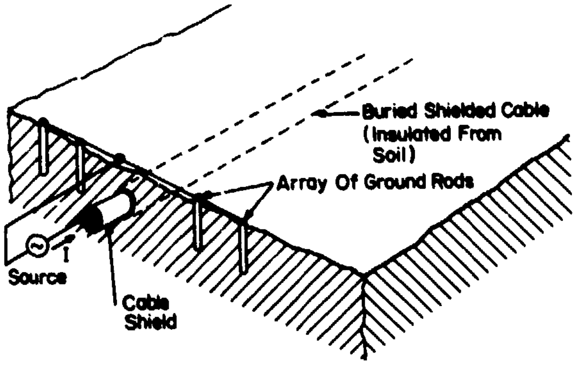


Coaxial cylinders can also be formed from the two outer shields of a double-shielded cable to drive a current in the inner shield. This scheme, as illustrated, is very efficient in terms of the pulse driver requirements because only the current in the inner shield must be simulated and this current is often smaller than the total cable current induced by an incident EM wave. Whether or not this driving technique can be used depends on the characteristics of the shield system and the inner shield current waveform. To apply this technique or any direct injection technique, the designer must have prior knowledge of the inner shield current waveform (from either test or analysis) to determine the coupling between the incident EM wave and the inner shield.



SHIELD AS A COUPLING STRUCTURE FOR CURRENT INJECTION

A second variation of the coaxial cylinder coupling geometry, as illustrated, also makes use of the natural environment of the cable system. In this case, the shielded cable is buried in the soil and is insulated from the soil by its plastic jacket. Since the metal shield, plastic jacket, and soil form a natural coaxial geometry, they can be used as the coupling structure for producing current in the shield. If this driving scheme is to be effective, the insulating jacket on the cable must be free of penetrating cuts or abrasions throughout the length of the cable that is to be driven. It will also be necessary to establish a low impedance connection to the soil at the driving point so that an acceptable fraction of the source voltage is applied to the transmission line. Finally, the wave propagating on the transmission line must not be severely attenuated by the soil return path so that it is dissipated before a sufficient length of cable has been excited.

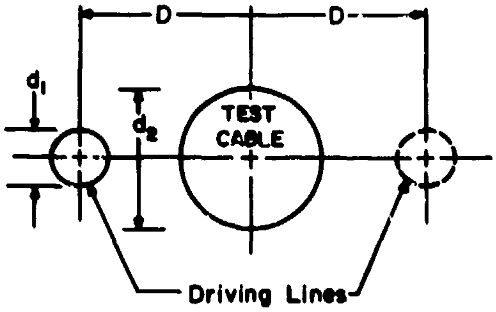


GROUND RETURN AS A COUPLING STRUCTURE FOR CURRENT INJECTION

In practice, the required bandwidth of the current pulse spectrum and the voltage limitations of the terminating resistor and capacitor bank require that the characteristic impedance be made as small as possible. The lowest characteristic impedances are available in coaxial transmission lines; however, it is difficult to construct such a line (except in those cases discussed above where the natural geometry of the system can be used), if the test cable is more than about one hundred feet long, particularly if the outer shield of the test cable is not insulated for high voltages. In spite of its very desirable electrical features (i.e., low characteristic impedance and uniform current distribution), this method of forming the transmission line has limited application because of the mechanical problems of drawing long cables through pipes and providing high-voltage insulation between the cable and the pipe.

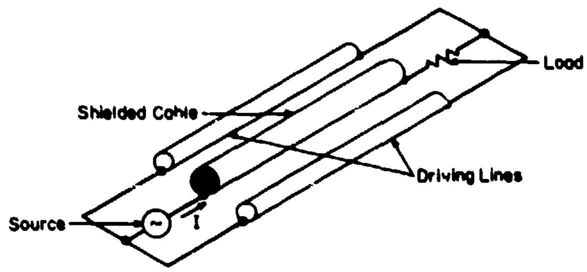
An alternative to the coaxial line is the parallel-wire transmission line. The characteristic impedance of a parallel-wire line with unequal diameters as illustrated in the figure is:

$$Z_o = \frac{120}{(\epsilon_r)^{1/2}} \cosh^{-1}\left[\frac{4D^2 - d_1^2 - d_2^2}{2\, d_1 d_2}\right]$$



PARALLEL-WIRE TRANSMISSION LINES WITH UNEQUAL WIRE DIAMETERS

where D is the wire spacing, d_1 and d_2 are the wire diameters, and ϵ_r is the dielectric constant of the insulating medium. When allowance is made for high-voltage insulation, it is difficult with a single driving conductor to obtain characteristic impedances of less than 100 ohms. It is possible to reduce this impedance by nearly 50 percent, however, by using two conductors in parallel (as illustrated by the second conductor indicated by the dashed line in the following figure) to drive the test cable.

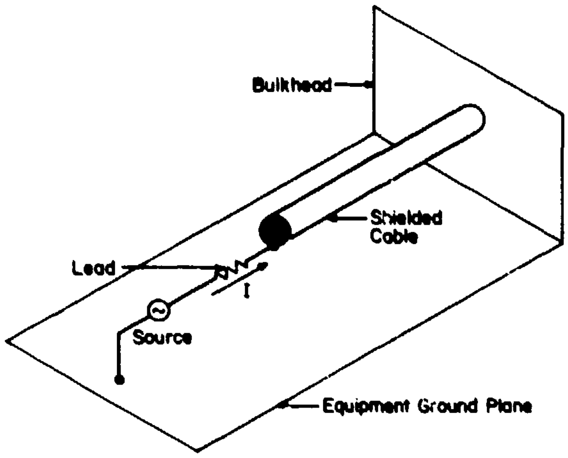


PARALLEL DRIVING LINES AS A COUPLING STRUCTURE FOR CURRENT INJECTION

This arrangement also produces a more uniform distribution of the current in the outer shield of the test cable and reduces magnetic coupling to the core of the cable.

It is fairly easy to construct a long, low-impedance uniform test line using the parallel-wire configuration as illustrated here. High voltage lines are used to drive the test cable, and the test cable is used as the low-voltage return for the parallel wire line.

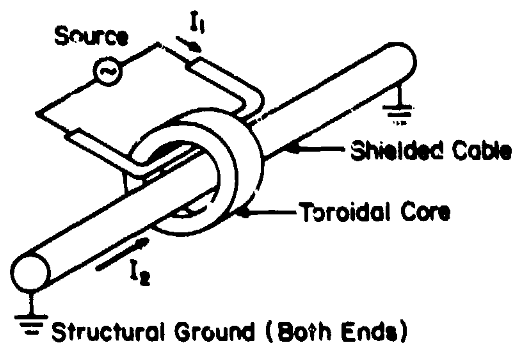
A variation of the parallel-wire driving structure is useful for driving shielded cables with insulating jackets that are routed along a metal structure or laid in metal cable trays as illustrated.



EQUIPMENT GROUND AS A COUPLING STRUCTURE FOR CURRENT INJECTION

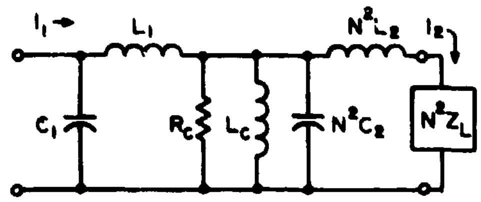
The configuration has the characteristics of a symmetrical two-wire transmission line, since the ground plane can be replaced by an image conductor to form a two-conductor line having twice the characteristic impedance of the conductor and ground plane. This method of driving the shield is limited to applications where one end of the shield can be removed from the ground and connected to the energy source. In cases where simulation is required, it has the advantage that much of the pulse shaping is accomplished by the system structure, if the cable length and terminations are preserved.

If both ends of the cable shield are grounded to the structure in the system, and it is desired to preserve this transmission line configuration so that the geometry of the system will shape the current waveform, then the current may be injected by means of a current transformer constructed as illustrated in the following figure. The toroidal core can be split and clamped around the cable without distrubing the cable system.



CURRENT TRANSFORMER TO INJECT CURRENT ON A GROUNDED CABLE SHIELD

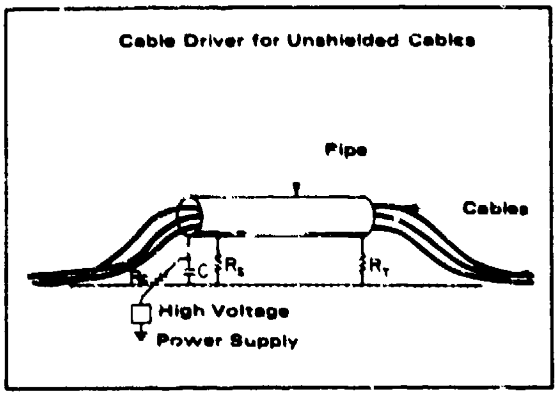
The equivalent circuit of an N turn primary and single-turn secondary current transformer referenced to the primary is illustrated below. L1 and C_1 are the primary inductance and stray capacitance, R_c is the core loss resistance, L_c is the core leakage inductance, and L_2 and C_2 are the secondary inductance and stray capacitance.



EQUIVALENT CIRCUIT OF A TOROIDAL CORE CURRENT TRANSFORMER

Unshielded Cable Driving Techniques

The current transformer concept can also be used for driving individual or multiple unshielded wires. For unshielded cables, signals can also be injected simultaneously on all or individual wires by small capacitance as is indicated here. The cables are placed within a conducting cylinder that forms the drive side of each coupling capacitance. This scheme is an easy method to loosly couple energy to multiple wires without disrupting the normal circuit operation.



This approach to simulating voltage injection into unshielded cable bundles induces currents that simulate the common-mode current that was measured in a low-level test of the complete system and assumes the currents will distribute properly between the individual wires in the bundle. Although this approach is useful, particularly in the early stages of the system test, the results can be misleading. Interconnecting cables between equipment cabinets within a shielded enclosure are often excited by currents conducted along a cable or along some of the wires in the cable rather than by coupling to ambient EM fields. Consequently, the current in the individual wires is not necessarily determined by specifying the total current in the bundle and the individual wire terminations. For example, assume that a bundle with many conductors contains some conductors that are connected to circuits outside a shielded enclosure as well as to some that are interconnecting conductors for circuits totally within a shielded enclosure. Thus, the interconnecting bundle contains some conductors that are tightly coupled to the conductors in the external bundle. In such a case, it is probable that the excitation of the interconnecting bundle will be dominated by the signal on those wires from outside the shield and that the signal on the remainder of the conductors in the interconnecting bundle will result primarily from mutual coupling among the conductors. A bulk common-mode current injected into the interconnecting wire-bundle, therefore, will not produce the same system response in the equipment as the bulk current produced by the EMP environment.

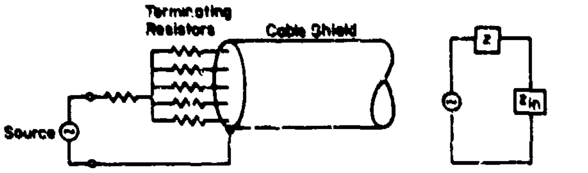
In the case described above, the bulk-current injection would produce a much more effective simulation if it were made on the wire bundle entering from the external equipment. If it were not possible to inject the test pulse into the external conductors (for example, only the interconnecting bundle was easily accessible), each conductor in the interconnecting bundle might have to be driven individually. Usually, it is possible to rank the importance of the individual wires in terms of the susceptibility of the circuits served by the wire so that precise simulation is not necessary for every wire in the bundle.

Direct Injection on Signal-Carrying Conductors

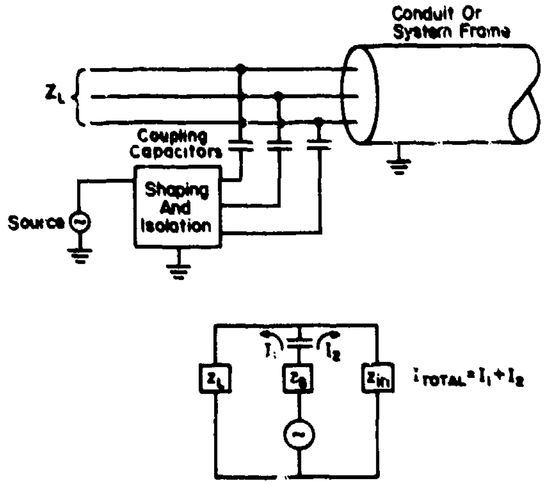
The injection of the test signal into signal-carrying conductors, such as the core conductors of a shielded cable or the conductors of an unshielded cable or equipment terminals, usually requires a more carefully designed experiment than the injection of test signals into cable shields. Most tests in which the cable shield is driven utilize loose coupling between the driving source and the signal-carrying conductors because the shield usually provides more than 20 dB of isolation. Because of this loose coupling, the system impedances that affect the responses of the signal-carrying conductors are not significantly affected by the driving system and the responses essentially occur from natural excitation of the cable shield. When the test signal is injected directly however, the driving source and coupling-system impedances may alter the system response. Thus, additional effort may be required to evaluate the effect of these differences - that is, to determine the system response had the effects of the injection system not been present or to place bounds on the possible effects of the injection system.

Direct injection of test signals on signal-carrying conductors of shielded cables also requires a more comprehensive understanding of the interaction of the system with the electromagnetic pulse, inasmuch as the designer must be able to determine (either through analysis or experiment) the EMP signal that couples to and penetrates through the cable shield. For unshielded cables, antennas, and power lines directly exposed to electromagnetic radiation, however, the problem of specifying the pulse shape may be no more difficult than that of specifying the pulse shape for shielded cables. The quality of simulation required (or the analysis needed to justify the quality of simulation used) may also be considerably greater for direct injection into signal-carrying conductors because all pulse shaping must be accomplished by the driving source and coupling network. The designer cannot take advantage of the pulse-shaping and loose coupling characteristics of the shield as is the case when testing with injection of currents on cable shields.

Direct injection (hard wire connection) of common mode voltages into individual wires of a cable bundle or directly at the equipment terminals can be achieved through an impedance matrix as shown in the following figure.



IMPEDANCE MATRIX USED TO INJECT COMMON-MODE VOLTAGES



CAPACITORS USED TO INJECT COMMON-MODE VOLTAGES

The impedance matrix should simulate the normal impedances between the conductors and between the conductors and the cable shield or system ground.

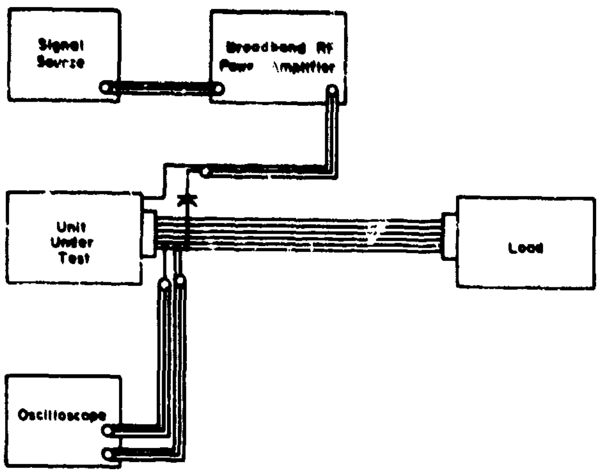
This method of driving cable conductors is perhaps the most straightforward and commonly used of all the direct injection methods. It can also be used with unshielded cables that are routed along a metal structure (such as an aircraft wing) or are placed in metal cable trays. With unshielded cables of this type, the current is driven against the metal structure or trays rather than against the shield. One disadvantage of this method is that the cable being driven must be disconnected at one end; hence, the system may not be operating in its normal state during the test.

When one end of the cable is not accessible, another injection method must be used. Such cases arise where disconnecting the cable precludes operating the system in its normal mode - for example, disconnecting the main power leads to inject a signal on these leads precludes operating the system from power supplied through the leads. In these cases, it may be necessary to accept some compromise in the quality of the simulation to perform tests economically. One approach that can be used under certain conditions is illustrated in the figure.

At some suitable junction in the cable system, the energy source is capacitively coupled to the conductors and permitted to drive them with respect to the local ground or chassis. As illustrated in the figure, however, the current injected at this point is divided into two parts, one flowing in each direction from the injection point. Because this method of distributing the current differs radically from the current distribution that would have resulted from the EMP excitation of the system, some care is required in designing a valid test using this approach.

Direct injection at the equipment terminals is often accomplished by driving terminal pairs or between a single terminal and ground. While this form of direct injection does not provide the normal excitation (all cable entries) of the system, it is useful to determine damage thresholds of individual circuits within the equipment or subsystem. This form of testing is applied principally to individual module testing for compliance to an EMP specification.

The basic test set-up is presented in the following figure.



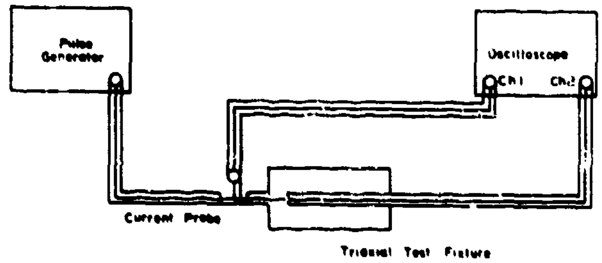
BASIC DIRECT INJECTION TEST SETUP

For this approach, the signal source is determined based on the test objectives. For damage threshold testing, the system should be operating at the time of signal injection. The unit should be terminated in the actual load or a simulated load with the same response. For upset testing, the unit under test must be operational. In this case, the terminal loading being the actual load is desirable. If this is not practical, a simulated load may be used. The simulated load should have the same frequency response characteristics of the actual load.

CW measurements are also commonly used to assess system response to an EMP pulse. Usually these tests consist of measuring the pin-to-pin, or pin-to-case impedance as a function of frequency. The basic equipment for this type of test consists of a swept frequency generator and a network analyzer that automatically displays both amplitude and phase of the equipment terminal impedance.

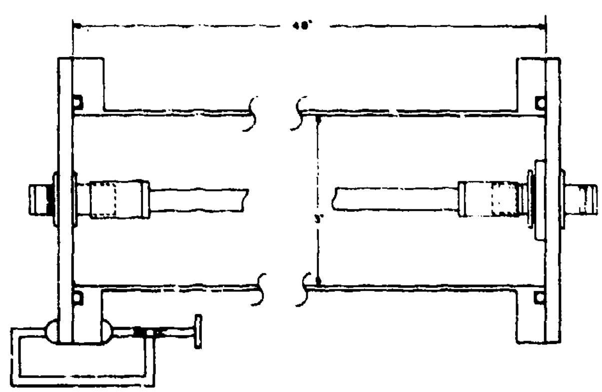
<u>Laboratory Component Tests - Cable Transfer Impedance</u>

Transfer impedance measurements of shielded cables can be made in the laboratory. The shield is driven as a part of a transmission line, in this case a coaxial line. The storage capacitor is discharged through a resistor that equals the coaxial line impedance. The resistor absorbs the reflection that occurs at the far end of the shorted line. The charging capacitor and resistor determine the pulse decay time which is many microseconds. Thus, the driving current contains considerable low frequencies. The reflections that occur in the line are at frequencies that are higher than those of interest for solid shield cables. If braid shields were being measured, the shorted end of the line could be terminated to give a fast, clean pulse rise. A typical test setup is shown in the following figure for testing in the time domain.



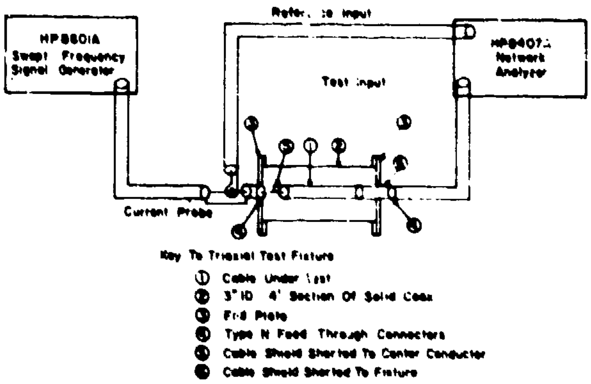
SURFACE TRANSFER IMPEDANCE MEASURENT TEST SETUP FOR TIME DOMAIN

The basic construction of the triaxial tester for use in these tests is illustrated in the following figure. This configuration allows for a 1 meter length of cable to be tested which is the normalized length for reporting the data.



TRIAXIAL TEST FIXTURE FOR SURFACE TRANSFER IMPEDANCE MEASUREMENTS

Transfer impedance measurements can also be made on a CW basis. If CW is utilized, measurements of both amplitude and phase (complex impedance) must be made at each frequency of interest. A typical test setup for frequency domain measurement of a cable transfer impedance is illustrated in the following figure. In this measurement setup, a swept frequency generator and a network analyzer are utilized to circumvent the tedious point-by-point measurement task, although point-by-point measurement is a suitable alternative.

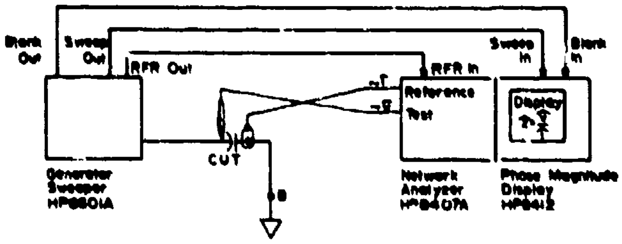


TEST SETUP FOR SURFACE TRANSFER IMPEDANCE MEASUREMENTS

Passive Components

Two basic measurements are required on passive circuit components. First, the frequency response of the device must be known in order to model the devices for use in an analytical program. Second, the pulse breakdown characteristics of the device must be determined, that is the damage threshold.

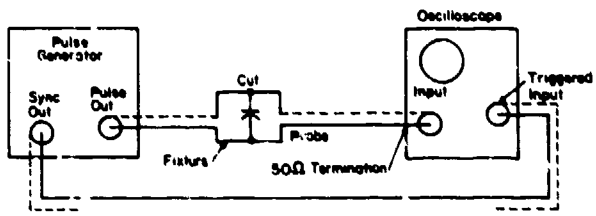
The basic test setup for determining the frequency response of the device is illustrated in the following figure. The figure shows a capacitor under test but the same setup applied for most two terminal devices. Devices such as transformers (four terminal devices) require a modification of the setup.



SWEPT IMPEDANCE MEASUREMENT SETUP

The lead length connecting the source to the unit under test and the ground lead must be as short as possible so the lead inductance does not cause a significant perturbation of the frequency response.

To determine the pulse damage threshold of a device requires a pulse source with a fast rise time (equivalent to the rate of rise of the EMP transient). The pulse breakdown rating is usually several times the dc rating, so the pulse source should be capable of several kV output voltage. As in the case of the frequency response measurements, the leads on the unit under test must be kept very short (< ¼ inch) or the lead inductance will seriously impair the measurement. A special low inductance fixture is usually required.

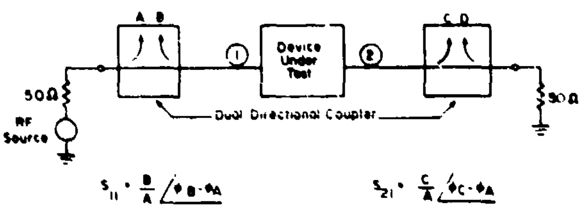


PULSE TESTING OF CAPACITOR

The test procedure for the low-level frequency domain characterization of filters is the measurement of the S parameters. Based on the typical EMP-induced transients, these measurements should be performed at least over a very wide frequency range. This procedure considers development of scattering parameters by reflectivity measurements. An alternate procedure considers development of scattering parameters by impedance measurement techniques. Both procedures are applicable over the frequency range of interest, generally from 1 kHz to 100 MHz. However, the reflectivity measurements are more appropriate for frequencies above 100 kHz, partly because of equipment availability. Above 10 MHz, the reflectivity measurements can be more accurate and simpler than impedance measurements.

The S parameters reduce to voltage reflection and transmission coefficients when characteristic impedance terminations and source impedances are employed. Therefore, the S-parameters can be easily measured with commercially available test equipment.

One of the standard circuits for measuring S-parameters is shown in the following figure.



STANDARD CIRCUIT FOR MEASURING S-PARAMETERS

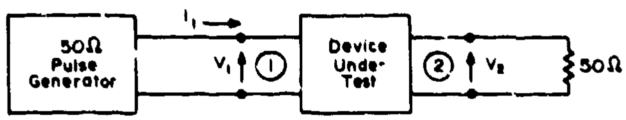
The setup shown, that is, the generator connected to port 1, provides the S_{11} and S_{21} measurements. Connecting the generator to port 2 provides the S_{12} and S_{22} parameters through the relationships

$$S_{12} = \frac{B}{D} \quad \phi_E - \phi_D$$

$$S_{22} = \frac{C}{D} \quad \phi_C - \phi_D$$

By knowing the "S" parameters, the filter response for any waveform and input and output terminations can be determined analytically. This is essential since the loads are not purely resistive over the entire frequency range of interest.

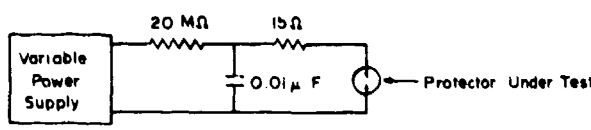
To obtain the damage threshold for filters, the concept illustrated is employed for low or high pass filters. This concept requires a high voltage pulser with a fast rise time. It is not adequate for narrow band pass filters since the energy spectral density in the filter pass band is generally insufficient to produce filter breakdown. In this case, a high level pulsed RF source would be more appropriate. The test should be conducted for both ports of the filter.



PULSE TEST PROCEDURE FOR FILTERS

Terminal Protective Devices

Two sets of measurements are required to characterize the performance of terminal protective devices for EMP application. First, a set of quasi-static measurements is required to evaluate the static breakdown, extinguishing voltage, follow current, etc., of the device. A typical test setup is shown for measurement of the static breakdown voltage.

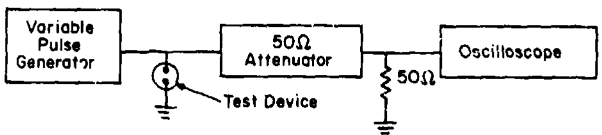


TYPICAL CIRCUIT FOR MEASUREMENT OF V_{SB}

The capacitor is incorporated so the power supply is not damaged due to the short circuit current when the device fires. To determine the extinguishing voltage, a second regulated, protected source is required.

The second set of measurements relates to the device capability to respond to a transient with a fast rate of rise. In this case, a high voltage pulse generator with a variable rate of rise is required to determine the response time as shown.

Since a short duration pulse is utilized in this approach, the short circuit current is usually not a problem if the pulser is protected (current limited). Again, since a transient is involved, lead lengths must be kept short to eliminate the effects of lead inductance. Usually low inductance fixtures must be fabricated to assure valid test data.



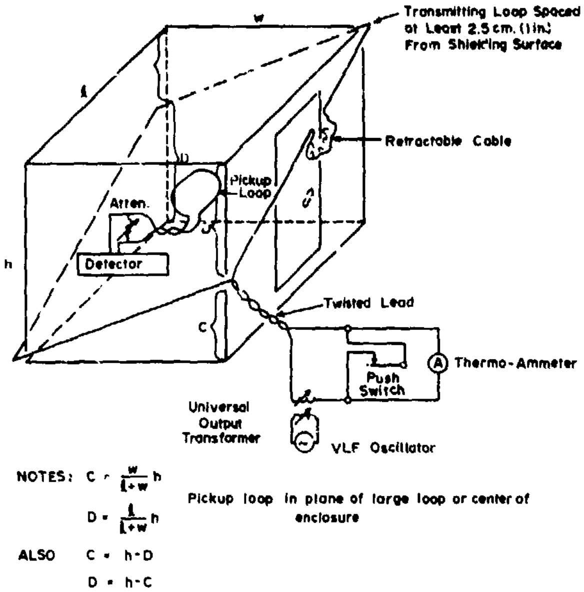
TYPICAL CIRCUIT CONFIGURATION FOR MEASURING V_{PB}

A frequency response measurement is also required to determine the normal insertion loss of these devices. This measurement can be performed by the same approach as for passive devices.

Shielding Measurements

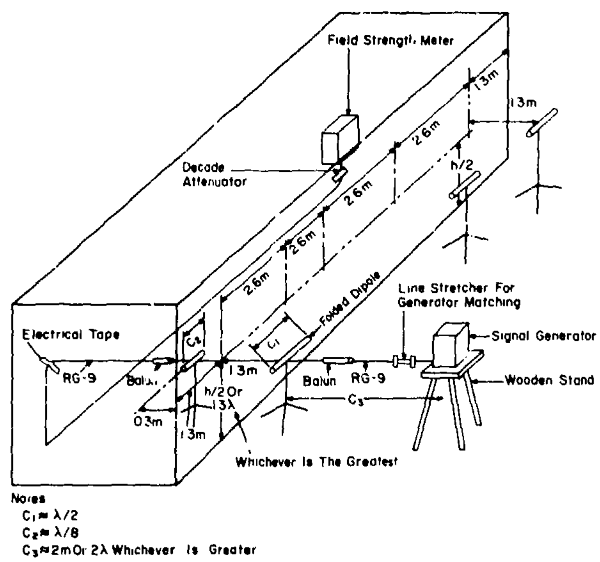
Shielding measurements are required on a variety of enclosure sizes. Ideally, the enclosure would be exposed to plane wave fields in one of the simulator types previously discussed. Many times, however, it is desirable to evaluate the enclosure performance in the laboratory. Standard measurement procedures are available.

Room size enclosures (a few meters on a side) can be evaluated in the low frequency (< 1 MHz) region by the large loop test setup illustrated. This procedure, while not providing the shielding effectiveness for plane waves, does excite currents on all faces of the enclosure. This would locate such features as poor seams, door gasketing, etc. This approach can provide for good quality control if the enclosure shielding effectiveness has been certified by plane wave illumination.



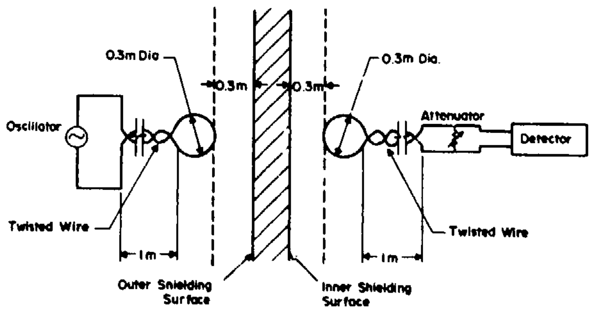
LARGE-LOOP TEST SETUP

To determine the high frequency response, the test setup illustrated is utilized in the UHF (30-300 MHz) band. This approach tests the enclosure on a piecemeal basis and again is good for a relative measurement in the laboratory. It will reveal flaws in the enclosure.



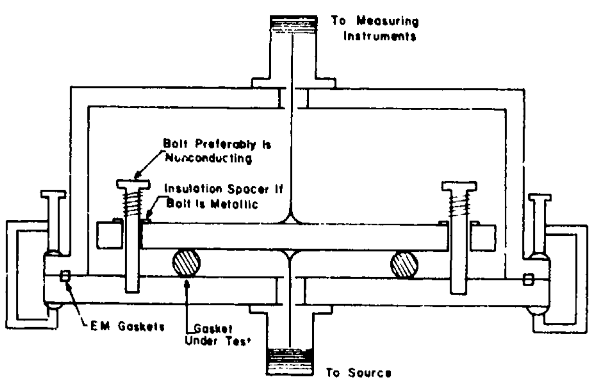
UHF MEASUREMENT SETUP

Measurements of seam performance, door gasketing performance, or locating apertures, etc., can also be performed using the small loop concept illustrated. This concept measures the shielding effectiveness in the local area of the loop. By probing the enclosure, flaws can be located.



SMALL-LOOP TEST SETUP

The independent determination of gasketing performance is often desirable. To relate the relative performance of various types of gasket materials, a standard fixture, as illustrated, is required.



SCHEMATIC CROSS-SECTION OF TEST FIXTURE USING BACKSHELL SHIELD

Using a standard fixture, the gasket performance can be documented on a normalized transfer impedance (Z_{TN}) given by

$$Z_{TN} = Z_T(\omega)\ \ell \quad \text{ohm-meters}$$

where

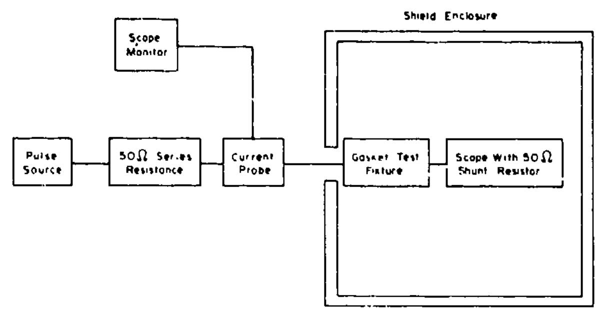
$$Z_T(\omega) = \frac{V_s(\omega)}{I_s(\omega)} \quad \text{ohms}$$

$V_s(\omega)$ = coupled voltage across the inner surface of the gasket

$I_s(\omega)$ = sheath current across the external part of the gasket

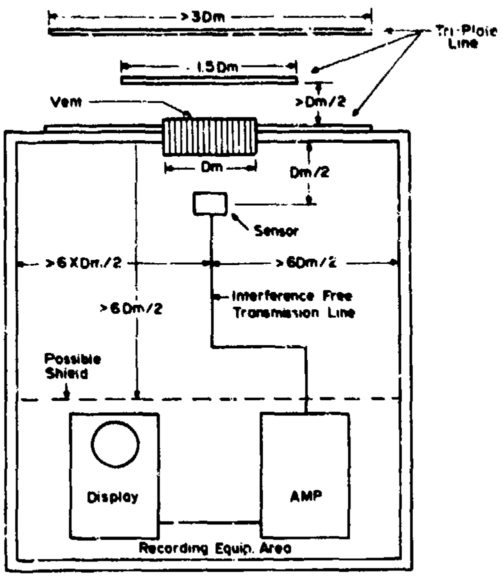
ℓ = gasket length.

The test setup for performing these measurements requires only standard laboratory instrumentation, plus a well shielded enclosure as illustrated. The pulse source rise time and duration must be selected to provide a spectral content at least as wide as the EMP transient.



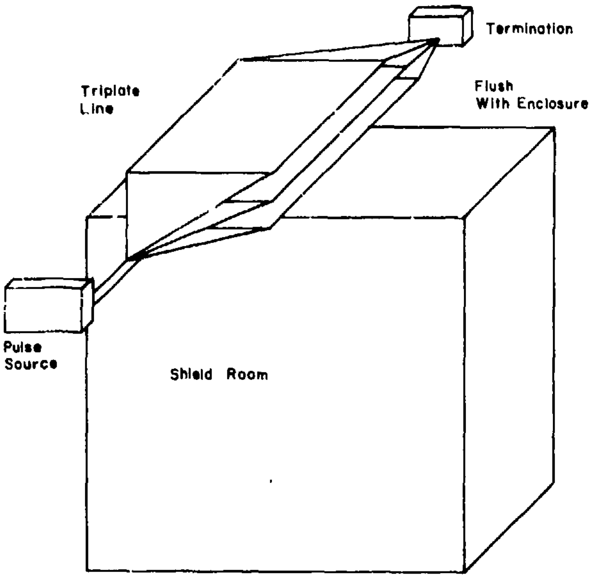
PREFERRED TEST SETUP FOR GASKETS

Evaluating the shielding effectiveness of vents (honeycomb, screens, etc.) is another important aspect of shielding measurements. Again, a standard test fixture is required to obtain relative performance, as illustrated.



SUGGESTED TEXT FIXTURE

A tri-plate transmission line is used as the exciting source in determining the shielding effectiveness for plane waves--as is illustrated below.



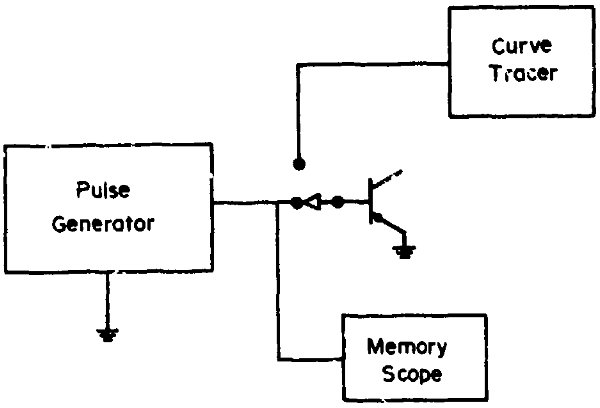
SUGGESTED EXTERNAL TEST

A tri-plate line is utilized for greater uniformity of the incident field. Utilizing a tri-plate structure with a narrower center plate reduces fringing of the field at the edges of the line providing a plane wave over the vent being tested.

The setup shown is for the E-field polarization normal to the plane of the vent. For complete assessment of the gasket performance, the other orthogonal polarizations of the E-field should be used. This can be accomplished by the same concept with alternate configurations of the illuminating structure.

Semiconductor Device Damage Threshold Testing

Discrete semiconductor device damage thresholds are normally obtained for the b-e junction as discussed in Section IV. The normal procedure is to test the device to destruction using a laboratory pulse generator as a source, as illustrated.



SEMICONDUCTOR TEST SETUP

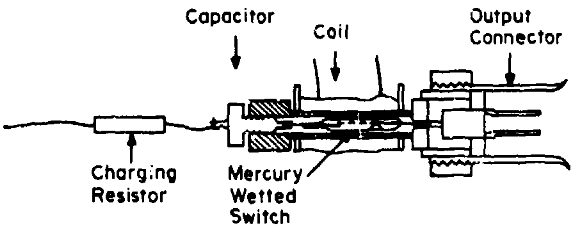
The pulse source should have both a variable amplitude (for step testing) and pulse width so the time dependence (slope of the damage curve) can be obtained. The setup shown is manual operation. A similar setup can be used for evaluating each terminal of an IC.

Current limited pulsers can also be used to advantage. Such pulsers could have the short circuit current limited to a value such that when the semiconductor goes into thermal second breakdown, the current through the device is below that required for destruction of the junction.

If a large number of devices are to be tested, automated test procedures should be considered. Automated test setups for making these measurements are available at the Air Force Weapons Laboratory.

7.6 DIMENSIONAL SCALE MODELING TECHNIQUES

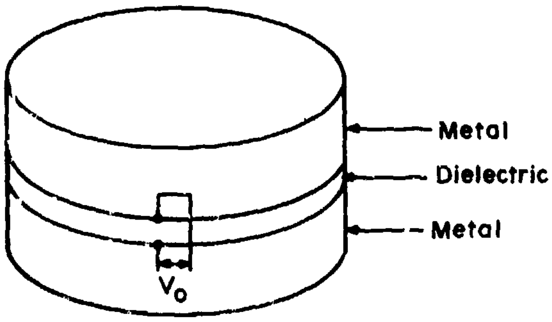
EM scale modeling of systems to determine their responses to transient EM fields is generally limited in the choice of the scaling parameter by available transient energy sources. When these sources are in the form of switch discharged capacitors, they can achieve transient rise times of the order of 10^{-10} seconds through the use of mercury-wetted reed type relays with contacts under gas pressure in a coaxial geometry as illustrated. The capacitor is charged through the charging resistor and discharged through the coaxial geometry, which is formed by the switch and switch housing. The capacitor can be replaced with coaxial cable to form square output waveforms and the entire system can be designed to operate at the coaxial cable impedance. Energy sources of the type illustrated can be constructed to produce output voltages from 1 to 4 kV. They can be easily converted to repetitive pulses at ac power frequencies, and being single-ended, can drive any scaled field simulator that has a single-ended input (for example, a monopole or parallel-plate transmission line). By utilizing a balun, a long dipole can be used to radiate EM energy at any incidence angle or polarization.



SCALE-MODELING TRANSIENT ENERGY SOURCE

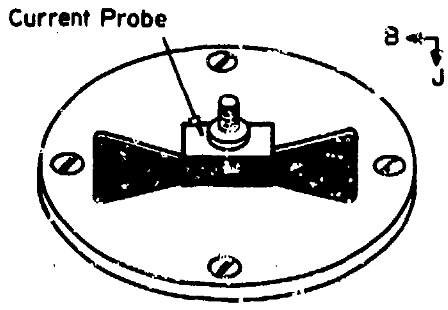
The system for receiving and recording scale model measurement waveforms consists of sensors, a sampling oscilloscope operated in the triggered mode, and an X-Y recorder to display the data. Sampling oscilloscopes can easily have rise time capabilities of 3×10^{-11} seconds. To trigger operation from a repetitive energy source, however, it is necessary to use a delay line in the signal channel to compensate for sweep delays in the oscilloscope. This delay line tends to limit the signal channel rise time to no better than 10^{-10} seconds.

Electric and magnetic fields and the currents on conductors and conducting surfaces of the models can be measured. Small capacitive probe antennas utilizing anodize as a dielectric such as that illustrated can be used to sense the electric field that is normal to a conducting surface. Such a sensor can have sufficient capacitance to drive a coaxial cable with a low frequency limit that is adequate for observing transient pulse widths of 20 ns. The sensor rise time can be less than 10^{-11} seconds.



ELECTRIC FIELD SENSOR

Commercially available probes for measuring current density, J, or magnetic field, B, with a small slot antenna, as illustrated in the following figure, have approximately 3×10^{-10} second rise time so the current measuring capability is the limiting factor in setting the scaling factor if the electromagnetic transient rise time is to be observed. Special small loops can be constructed to improve the current and magnetic field measurement rise time; however, they generally have decreased sensitivity, which limits their application in scale model measurements.



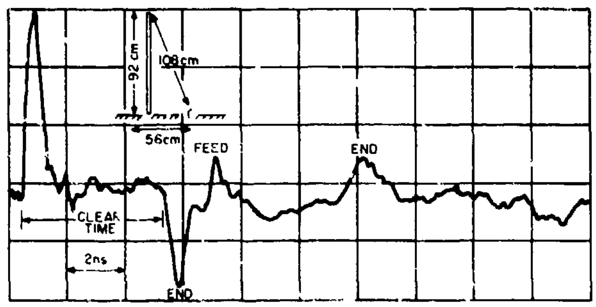
MAGNETIC FIELD SENSOR

EM scale modeling of a system entails a scaled construction of the conductive and EM features of that system. The various parts of the system are duplicated in miniature. However, only those features that define the EM characteristics need be properly scaled. Generally, only those features that can have a first-order effect on the interaction of the system with the incident wave must be reproduced in the scale model. Features that have little effect, such as internal conductors and small apertures, do not need to be reproduced; however, if they can be easily scaled, they may improve the fidelity of the model. EM scale modeling thus is used to study the gross features of the system that control the total system response. Secondary energy coupling effects that occur within the system are usually not readily amenable to scaling.

In the construction of EM scale models, EM planes of symmetry can sometimes be utilized to reduce the model size and to provide shielding for the instrumentation. Generally, symmetry planes can only be utilized when modeling simulators and very special systems such as missiles or aircraft. The conductive materials used in scale models are commonly copper sheet and wire, tubing, and mesh. Earth is modeled by adding salt to real earth to increase conductivity. Dielectric materials are generally modeled with plastics or wood.

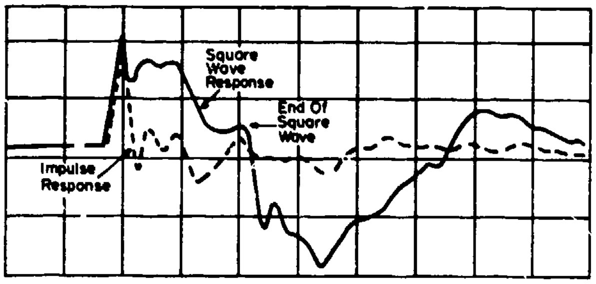
The scale factor selected for modeling an EM response (or coupling) problem depends on the purpose of the modeling test. If the response caused by a 10^{-8} to 10^{-9} second transient rise time is to be investigated, then a scale factor of less than 50 will be required if the scale model measurements are to observe this rise time. For large systems, it is desirable to use larger scale factors. Fortunately, such factors can be used because the rise time response of systems is approximately the excitation rise time convolved with the time corresponding to the dimensions of the system that are parallel to the incident field polarization. When using scale factors that give a faster response than the measurement system response rise time, the data should be checked to verify that the rise time is not limited by the measurement system.

When scale modeling transient EM coupling, it is possible to scale the EMP transient waveform and thus obtain response data that can be directly interpreted for the transient response of the full-scale system or for direct injection waveforms. In scale modeling EM field simulators or an entire system to determine the response of the components of the system, it is often convenient to use a short (or impulse-like) transient so that clear times and system responses can be separated. The following figure shows an example of this approach for a modeled monopole. Both the clear time and the multiple reflections on the monopole are directly observed and can be interpreted in travel times on the scaled model.



SCALE MODEL MEASUREMENT OF MONOPOLE REFLECTIONS

For observing the natural response of a system, it is preferable to use a long square wave with low frequency energies so that the system's natural resonances can be observed. The figure shows the response of a system to a short (impulse-like) transient and its response to a long, square-wave transient. The square-wave response is the integral of the impulse response and provides a direct measurement of the late-time (low frequency) response characteristics that are not easy to observe from the impulse response. With careful interpretation, almost any transient that excites the scaled system can be used to determine the response characteristics of that system.



SYSTEM IMPULSE AND SQUARE WAVE RESPONSE

7.7 SIMULATION FACILITIES

As an aid to the reader, brief descriptions of representative large scale EMP simulators are presented. Tables listing the basic characteristics of these and additional simulators is also provided.

Bounded Wave Simulators

ARES

The Advanced Research EMP Simulation facility is another bounded-wave transmission line simulator. ARES is a DNA facility located on Kirtland Air Force Base, New Mexico. The normal working volume is 40 m high, 30 m long, and 40 m wide. The transition sections are 76 m long. The line impedance is about 100 Ω.

The ARES pulser has a nominal charging voltage of 6 MV and energy storage of 50 kj. The peak output voltage is 45 Mv maximum. The energy source is a coaxial line storage element and a Van de Graaff generator power supply.

Field strengths on the order of approximately 110 kV/m are obtainable in the interaction volume. The pulse rise time is 6 nsec. The pulse width is variable from 100-500 nsec.

Housed beneath the facility is a shielded instrumentation room containing data recording and monitoring instrumentation. In addition to hardwire data links, a multichannel microwave data link is also available.

TRESTLE

The TRESTLE simulator is presently under construction at Kirtland Air Force Base, New Mexico. It is similar in appearance to ARES, but considerably larger. It will produce fields in excess of 50 kV/m over an interaction volume of approximately 80 l x 80 w x 75 h meters. The unique feature is a trestle for supporting the system under test in the center of the interaction volume.

ALECS

The AFWL-LASL Electromagnetic Calibration and Simulation facility is a bounded-wave transmission line simulator located south of the east-west runway at Kirtland Air Force Base, New Mexico. The working volume is 13 m high, 15 m long, and 24 m wide. The transition sections are 50 m long. The line impedance is 95 Ω.

The electromagnetic pulse is produced by discharging a graded capacitor stack into a short, 41-Ω coaxial transmission line (the peaking capacitor concept) and by discharging this series-combination into a transition structure that mates with the ground plane and parallel-wire top plane of the array. The 41-Ω coaxial line provides for the initial, nanosecond rise time of the wave, while the graded capacitor stack provides for the large amount of energy in the remaining portion of the wave. The pulse launching system is energized by a 2.2 Mv Van de Graaff generator prior to discharge. A triggered gap provides about one pulse per minute maximum.

The electromagnetic pulse that is delivered to the working volume has an approximate 1.7 Mv peak voltage, a rise time of approximately 6 nsec, and a 150-nsec decay (1/e). An experiment in the working volume will see a peak E-field of up to approximately 125 kV/m.

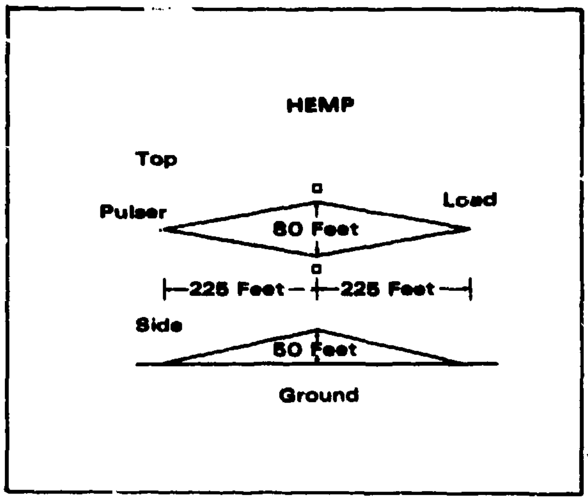
An underground, shielded room contains diagnostic instrumentation to implement the desired testing programs. Hardwire instrumentation links may be routed from the working volume of the simulator and the instrumentation room.

The ALECS instrumentation system consists of a ten-channel microwave-data system, a hardwire data system, a trace recording system, a timing and firing system, and attendant checkout and monitoring equipment. Most of the above systems are housed in the double-walled, welded-seam, sheet-steel shielded rooms that afford a 120-dB attenuation.

HEMP

HEMP is another bounded-wave transmission line simulator located at the White Sands Missile Range, New Mexico. This simulator does not have a working volume section. The two transition sections are each 68 m long. The line dimensions where the transition sections meet is 15 m high and 24 m wide. The line impedance is 135 Ω.

The transient source is a 10-stage LC generator that can provide voltages of 50 to 400 kV. At 400 kV, the peak electric field is 26 kV/m with a 3 nsec rise time. The transient decay time constant is about 300 nsec.

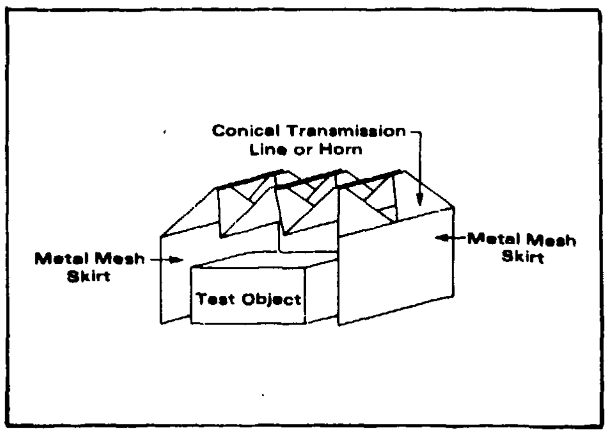


TEFS

The Transportable Electromagnetic Field Simulator is a bounded-wave transmission line simulator with multiple feeds that is designed to propagate a transient in the vertical downward direction. Five-hundred and seventy-six transition sections, each with a line impedance of 200 Ω are used. Four transitions are paralleled and driven from a 50 Ω cable.

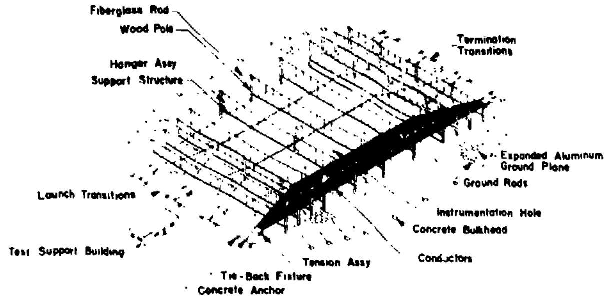
The cables (144 total) are commonly driven from a single switch and capacitor bank. The 144 sections can be configured in a variety of ways to illuminate an area of 40 x 40 meters. A field of 50 kV/m is provided with a 4 nsec rise time and a decay time constant of 350 nsec.

Versions of this simulator are available at the White Sands Missile Range, New Mexico and the Naval Surface Weapons Center/White Oak Laboratory test facility at Patuxent Naval Air Station.



SEIGE

The Simulator Induced EMP Ground Environment simulator is a bounded-wave transmission line simulator with multiple feeds or transition sections. This simulator is designed to test a buried system. The working volume is 3 m high and uses the earth for the lower plate. Four 80-Ω transitions drive the line. One version of this simulator is at the Air Force Weapons Laboratory, Kirtland Air Force Base, New Mexico.



Long Wire Dipole Simulators

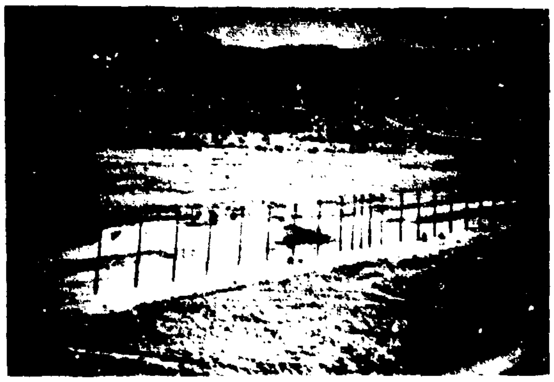
Sandia Long-Wire

One of the first EMP simulators constructed was the Sandia Long-Wire. This facility, located on Kirtland Air Force Base, New Mexico, was designed and constructed by the Sandia Corporation. This simulator is a horizontal dipole radiator with discrete lumped loading resistors.

The "long-wire" in the Sandia Long-Wire was actually a 6 inch diameter, aluminum irrigation pipe. Fifty sections of this pipe, each 20 feet long, were joined to form a 1,000 foot long-wire antenna. The antenna was supported by telephone poles, 40 feet above the ground.

Two 20 kV, dc-power supplies were connected to a pressurized parallel-plate switch. Each half of the antenna is charged by the 20 kV power supplies, with one side positive and the other negative. The center switch then would break down at a voltage level as determined by the gas pressure and gap separation. The output pulse is approximately that of a double exponential, with a rise time of less than 10 nsec, and a decay time (peak to 10 percent) of approximately one microsecond.

The test object (such as a missile) is positioned approximately 100 feet from the antenna and 10 to 20 feet above the ground. Under these circumstances, the system under test would see a peak E-field of approximately 1000 V/m.

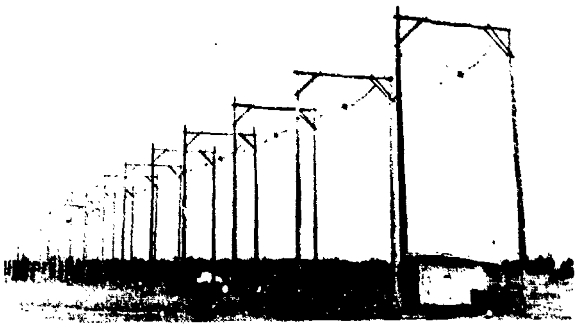


Martin-Marietta Long-Wire

The Martin Marietta Long-Wire facility is located at Orlando, Florida. This facility is similar to the Sandia Long-Wire. The antenna is 1000 feet long and 46 feet above the ground. It is powered by two 125 kV power supplies, and corona rings are used to minimize corona losses.

The spark gap uses a mixture of 6 percent oxygen and 94 percent nitrogen, under pressure (50 to 5000 psi). By varying the pressure, the spacing of the gap, the charging resistance, and the voltage, it is possible to vary the pulse repetition frequency, the amplitude of the radiated field, and the rise time of the pulse.

The pulse rise times can be varied from 5 to 30 nsec, and the pulse widths can be varied from 100 to 700 nsec. Maximum peak E-field intensity is 1100 V/m at a point 100 feet from the antenna. Pulse repetition rates of 10 per second are possible.



1000 Foot Long, Resistively Loaded Spark Gap Excited EMP Facility

Pulsed Radiating Simulator Facilities

TEMPS

The Transportable Electromagnetic Pulse Simulator (TEMPS) is a transportable biconic dipole radiator utilizing two back-to-back Marx generator pulsers. The radiator can be varied in length in 100 meter increments to a maximum of 300 meters. The dipole is resistively terminated to earth. The pulser has an adjustable rise time of 4 to 12 ns and energy content of 60 kj. The pulse shape is a double exponential with a cross over (zero crossing) at approximately 800 ns. The output field strength at 50 m from the antenna on center line is 60 kV/m maximum, although normal operation is slightly less than this value. The interaction space is + 25 m on a line parallel to the antenna at 50 m for good wave planarity. Included with the system is a complete instrumentation van and data recording system, computer aided handling and analysis system, and support facilities.

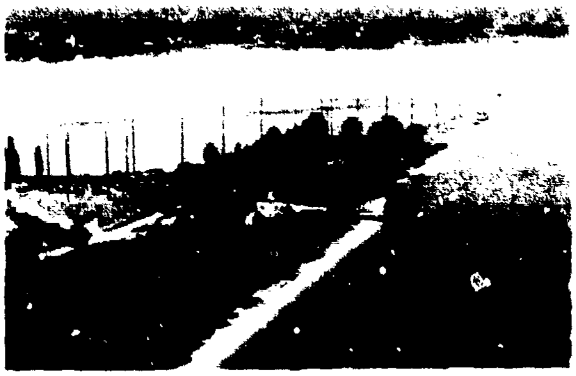
AESOP

The Army Electromagnetic Simulator Operations facility (AESOP) is an exact duplicate, in terms of EM characteristics, of the TEMPS. The only difference is the AESOP is a fixed installation. It is operated and maintained by the Harry Diamond Center and is located at the Woodbridge Research Facility, Woodbridge, Virginia.

EMPRESS

The Electromagnetic Pulse Radiation Environment Simulator for Ships (EMPRESS) facility is operated by the Naval Surface Weapons Center/White Oak Laboratory and is located at Solomons, Maryland. It is capable of generating both vertically and horizontally polarized EM environments. In the horizontal mode, a 1,300 foot biconic dipole configuration is utilized. Field strengths of 2.2 kV/m can be generated at 300 m. The interaction area, for good wave planarity is approximately 300 m parallel to the antenna axis at a distance of 300 m. The waveform is a double exponential. The pulser is a Marx generator with a 2.5 MV output.

In the vertical mode, an inverted conical monopole 100 feet high is the primary antenna. The monopole is top loaded by the 1,300 foot horizontal antenna to produce the low frequency content. The same pulser is used in both the vertical and horizontal configurations. The waveform approximates that from a surface burst with an amplitude of 4 kV/m at 300 m.



IITRI Crystal Lake Facility

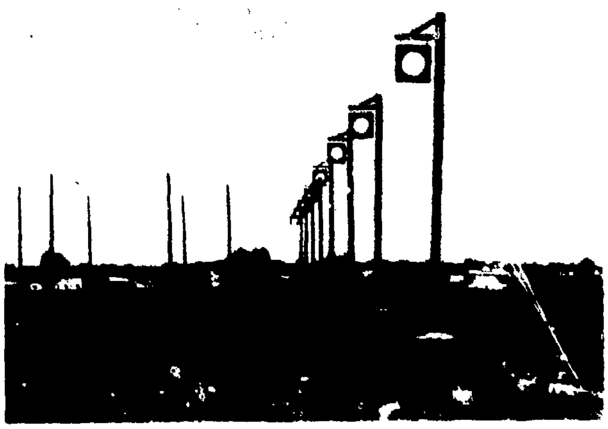
The IITRI facility is a scaled down version of the EMPRESS. It is a hybrid simulator that uses a 50 foot high vertical conic antenna and a 300 foot horizontal fringing line. This facility is located at Crystal Lake, Illinois on the IITRI property.

It is a sub-threat-level simulator with a peak radiated field up to 100 V/m single shot or 10 V/m on a repetitive basis at 60 pulses per second. Pulse rise time is approximately 10 nsec and the decay time approximately 10^3 to 10^5 nsec.

Essentially uniform fields are generated over an interaction area of 20 meters by 50 meters.

The conic antenna provides the high-frequency, short-time transient and the fringing quasi-static field from the horizontal wire provides the low frequency or late time transient. The two radiators are driven from a common source to provide a vertically polarized test field.

In the horizontal mode, the 300 foot fringing field line is converted to an earth terminated biconic dipole radiator.



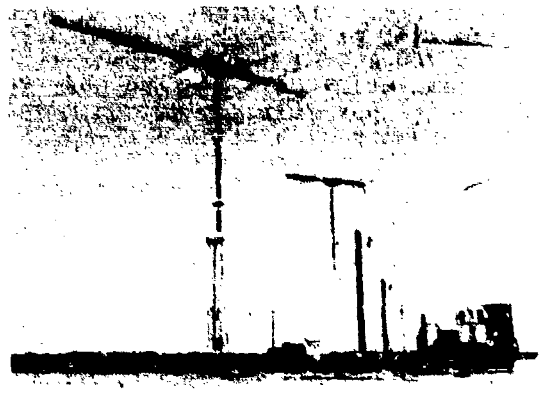
RES

The AFWL has a flyable Radiating Electromagnetic Simulator (RES) that is a biconic radiator. The horizontal version is 200 feet long and uses distributed resistive coating on the dipole elements to attenuate the antenna current to minimize reflections. The pulser is a Marx generator with an output voltage of 1.5 MV. The output waveform is a double exponential wave. The low frequency content is limited due to the short physical length of the antenna. Field strengths on the order of 1000 V/m at 500 m are possible. The unique characteristics of this facility are that it is easily transportable, and being airborne, can provide all angles of arrival of the incident pulse.

CW Radiating Facilities

WSMR Facilities

The U.S. Army at White Sands Missile Range, New Mexico, has a 10 to 200 MHz CW facility that uses log periodic antennas to obtain 1 V/m fields either vertically or horizontally polarized.



HDL

The Woodbridge Research Facility of the Harry Diamond Center at Woodbridge, Virginia operates and maintains a CW facility for coupling tests. The facility utilizes LPA antennas to provide approximately 0.5 V/m fields over a frequency range of 3-100 MHz. The illuminated test area is approximately 30 x 30 meters at a distance of 240 meters. The transmitter is a 200 watt power amplifier unit. Both vertical and horizontal fields polarizations are available.

Characteristics of Available Simulators

The following tables present data on the existing simulators. Indicated are the technical characteristics, the location, and the operational status.

NAME	LOCATION	WAVE-FORM	POL.	MAGNITUDE	INTERACTION VOLUME(ℓ,ω,h)	STATUS
				BOUNDED WAVE SIMULATORS		
ALECS	Kirtland AFB, NM	Exo	V	50 kV/m	30x30x10 m	Operational
ARES	Kirtland AFB, NM	Exo	V	70 kV/m	40x30x30 m	Operational
TRESTLE	Kirtland AFB, NM	Exo	V	>50 kV/m	80x80x75 m Modular up to	Under Construction
TEFS	WSMR, NM	Exo	V,H	65 kV/m	40x40x10 m	Operational
TEFS	NSWC/WOL, MD	Exo	V,H	50 kV/m	Modular	Operational
				LONG-WIRE DIPOLE SIMULATORS		
SANDIA LONG-WIRE	Kirtland AFB, NM	Exo	H	1 kV/m	~ 10x10x10 m (~ plane wave-front)	Unknown
MMO LONG-WIRE	Orlando, Fla.	Exo	H	1 kV/m	~ 10x10x10 m (~ plane wave-front)	Operational

NAME	LOCATION	WAVE-FORM	POL.	DIRECT WAVE MAGNITUDE/DISTANCE	INTERACTION AREA (~Pl. Wave)	ANGLE OF ARRIVAL	STATUS
				RADIATING WAVE SIMULATORS			
RES I	Portable, Kirtland AFB	Exo	H	1000 V/m @ 500 m	100 m	any	Deacti-vated
RES II	Portable, Kirtland AFB	Exo	V	1000 V/m @ 500 m	100 m	any	Deacti-vated
VPD	Kirtland AFB, NM	Exo	V	3 kV/m @ 200 m		grazing	Opera-tional
HPD	Kirtland AFB, NM	Exo	H	*50 kV/m @ 9 m HAC	Area directly below antenna (nonplan-ar)	normal	Opera-tional
HDL Biconic	HDL, Woodbridge, VA	Exo	H	15 kV/m @ 100 m	~ 200 m	10^{o}@200 m	Opera-tional
AESOP	HDL, Woodbridge, VA	Exo	H	50 kV/m @ 50 m	~ 200 m	10^{o}@200 m	Opera-tional
VEMPS	HDL, Woodbridge, VA	Exo	V	5 kV/m @ 25 m (0.25 MV pulser)	~ 100 m	grazing	Opera-tional
EMPRESS	NSWC Solomons, MD	Exo	H	2.2 kV/m @ 300 m (16 m HAG)	~ 300 m	8^{o}@300 m	Opera-tional

*Directly below antenna

(continued)

NAME	LOCATION	WAVE-FORM	POL.	DIRECT WAVE MAGNITUDE/DISTANCE	INTERACTION AREA (~Pl. Wave)	ANGLE OF ARRIVAL	STATUS
				RADIATING WAVE SIMULATORS			
EMPRESS	NSWC Solomons, MD	Surface	V	4 kV/m @ 300 m	~300 m	grazing	Operational
EMPSAC	NSWC/NATC, Patuxent, MD	Exo	H	8.5 kV/m @ 50 m	~25-50 m	17°@50 m	Operational
NAVES	NSWC/NATC, Patuxent, MD	Exo	V	~11 kV/m @ 50 m	~25-50 m	grazing	In Constuction
TEMPS	DNA, Transportable	Exo	H	50 kV/m @ 50 m 12.5 kV/m @ 200 m	~200 m	10°@200 m	Operational

7.8 TEST INSTRUMENTATION AND SET-UP

Up to this point, we have discussed testing from the standpoint of generating a simulated EMP environment. The next consideration is monitoring the environment and the diagnostic instrumentation for determining the effects of the environment on the system under test. The primary consideration will be the nature of the instrumentation and the basic problems usually encountered. It will not consider sensor design.

Testing implies experimental observation of the EMP environment and the system response; this means the measurement of transient electromagnetic fields, voltages, and currents. A complete measurement system should include:

1. A signal sensing system
2. A distribution system
3. A recording system
4. A test plan and data reduction and processing system to interpret the data
5. Calibration including overall response characteristics.

Each of these areas will be discussed in subsequent sub-sections.

Signal Sensing

Signal sensing involves both the measurement of the field environment and the induced voltages and currents within the system. Both sensors and probes are required to perform these measurement functions.

Sensors refer to the measurement of electromagnetic fields. These can be in free space to monitor simulator fields, within a conducting medium such as earth, or within equipment racks or buildings.

Probes refer to the measurement of a voltage or current. Current probes measure the total current flow in wires. Voltage probes can be single ended referenced to a common signal line or can be differential.

Sensors

The most commonly used broadband electromagnetic field sensors are electrically small antennas. Such antennas have the directivity characteristics of a small electric or magnetic dipole and an impedance which is predominantly reactive. Short antennas or specially designed capacitors are used for sensing electric fields. Small loops are used for sensing the magnetic fields. The open-circuit voltage and short-circuit current response of electrically short antennas are proportional to either the field or the time derivative of the field as indicated.

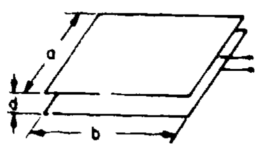
Output Response

	Voltage	Current
Probe	E	$\dot{E}$
Loop	$\dot{B}$	B

Electric Field Sensors

Essentially, any structure that can attain a potential with respect to ground or another portion of itself can be used as an electric field sensor. Any lossless small antenna has a capture or effective area, A_e, determined by the available power at the antenna terminals and the power per unit area of the EM wave. In practice, some incident power is lost because the antenna and its matching network dissipate energy and because the high antenna reactance generally limits the bandwidth over which an effective match can be made. Therefore, for an electrically small antenna, it is more useful to use the effective volume, S_e, defined as the ratio of energy stored in the antenna reactance to the energy per unit volume in the corresponding field of the incident wave.

One of the simplest versions of an electric field sensor is the capacitive plate antenna illustrated.



Rectangular Plates

$a \gg d$ $\qquad$ $b \gg d$

$\ell_e = \frac{V_o}{E} = d$ $\qquad$ $I_s = \epsilon \dot{E} A_e$

$A_e = ab$

$S_e = \ell_e A_e = abd$

The open circuit voltage (V_o) of this type sensor is the capacity weighted average of the potential that each plate would attain separately referenced to a common point. This open circuit voltage is given by:

$$V_o = E\,d$$

V_o = open circuit voltage

E = electric field intensity

d = plate separation.

The accepted definition of effective length (ℓ_e) of an antenna is given by:

$$\ell_e = \frac{V_o}{E} = d$$

The normal electric field and its associated displacement flux (D) induces a surface charge density (q) given by:

$$q = D = \epsilon E$$

ϵ = permittivity of the surrounding medium

The short circuit current (I_o) is the time rate of change of the charge density over the surface area of the plates (A):

$$I_o = \frac{d}{dt}\int_A q\,dA = \epsilon\,\dot{E}A = \epsilon\,\dot{E}ab$$

a = width of plate

b = length of plate

providing a, b >> d and neglecting fringing.

The "effective area" (A_e) is defined as the ratio of the short circuit current to the rate of change of displacement flux:

$$A_e = \frac{I_o}{\dot{D}} = ab$$

where

$$\dot{D} = \epsilon\,\dot{E}$$

which, for the case under consideration, is the physical area of one plate, neglecting fringing.

From Thevenin's theorem, the relation between the open circuit voltage and the short circuit current is the source impedance, which for this case is given by:

$$Z_a = \frac{V_o}{I_o} = X_{Ca}$$

Z_a = antenna impedance

X_{Ca} = capacitive reactance of antenna

which, for a sine wave, is:

$$X_{Ca} = \frac{1}{j\omega C_a}$$

and, therefore

$$C_a = \frac{\varepsilon ab}{d}$$

The peak energy per unit volume (W_f) in the incident field is:

$$W_f = \varepsilon|E|^2$$

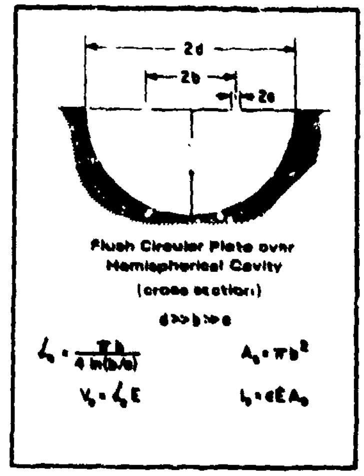
while the peak energy stored in the antenna (W_a) is:

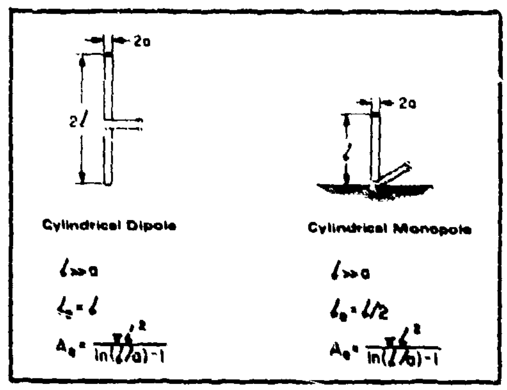
$$W_a = C_a V_o^{\,2}$$

The ratio is the "effective volume" (S_e) as defined previously:

$$S_e = \frac{W_a}{W_f} = \ell_e A_e$$

Similar expressions can be derived for other types of electric field sensors based on their geometry. The expressions for the flush antenna, dipole and monopole illustrated, are given in the following table.

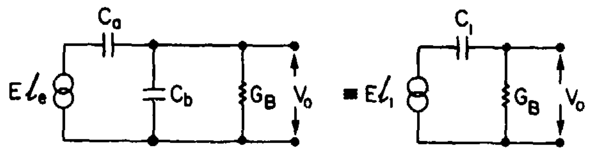




Type Of Antenna	ℓ_e	A_e	S_e	Capacitance	Conditions
Capacitive	$\left\lvert\frac{V_o}{E}\right\rvert$	$\left\lvert\frac{I_o}{\omega D}\right\rvert$	$\left\lvert\frac{V_o I_o}{\omega\varepsilon E}\right\rvert$	$\varepsilon A_e/\ell_e$	
Plate	d	ab	abd	$\varepsilon ab/d$	a>>d;b>>d
Flush Plate	$\frac{\pi b}{4\ln(\frac{b}{a})}$	πb^2	$\frac{\pi^2 b^3}{4\ln(\frac{b}{a})}$	$\varepsilon 4b\ln(\frac{b}{a})$	d >> b >> a
Dipole	ℓ	$\frac{\pi\ell^2}{\ln(\frac{\ell}{a})-1}$	$\frac{\pi\ell^2}{\ln(\frac{\ell}{a})-1}$	$\varepsilon\pi\frac{\ell}{\ln(\frac{\ell}{a})-1}$	ℓ>>a
Monopole	$\frac{\ell}{2}$	$\frac{\pi\ell^2}{\ln(\frac{\ell}{a})-1}$	$\frac{(\frac{\pi}{2})\ell^3}{\ln(\frac{\ell}{a})-1}$	$\varepsilon 2\pi\frac{\ell}{\ln(\frac{\ell}{a})-1}$	ℓ>>a

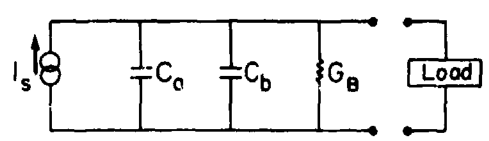
PROPERTIES OF SMALL ELECTRIC FIELD SENSORS

The voltage equivalent circuit of a probe antenna consists of a voltage source, V_o, an antenna capacitance, C_a, a base capacitance, C_b, and a loss element associated with its feed point and support structure. Normally the loss term, G_B, can be made small enough to be neglected. The base capacitance and antenna capacitance can be combined and a new effective length, ℓ_1, defined to give a simpler equivalent circuit. This equivalent length, ℓ_1, is the ratio of V_o/E including the voltage divider effect of C_a and C_b. Additional capacitance in parallel with C_b is often used to shorten the effective length or add attenuation to the sensor system.



$$C_a = \frac{\epsilon A_e}{\ell_e} \qquad C_1 = C_a + C_b \qquad \ell_1 = \frac{C_a \ell_e}{C_a + C_b}$$

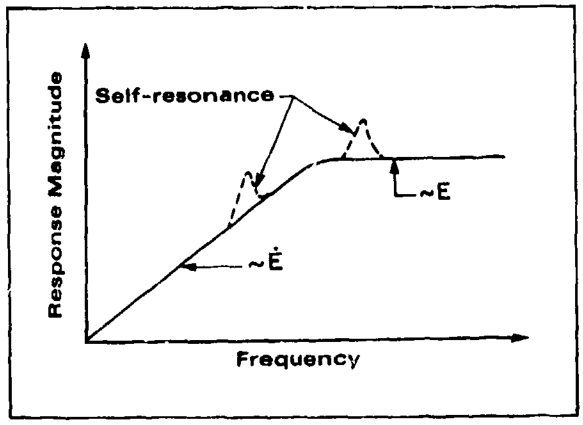
The capacitive probe antenna can also be described in terms of a current equivalent circuit as illustrated. When the load is a short-circuit, the current at the load is proportional to the time derivative of the incident electric field. This current is proportional to frequency, ω, and the magnitude of the electric field, giving an amplitude-frequency response that increases with frequency (time derivative).



$I_s = \epsilon \dot{E} A_e$ in time

$I_s = \epsilon \omega |E| A_e$ in frequency

When the load is finite but nonzero, for example the impedance of a coaxial cable, the probe antenna response is proportional to the time derivative of the electric field for frequencies below that at which the probe capacitive reactance is equal to the load impedance. For high frequencies, the response is proportional to the electric field.



So far, the probe antenna has been considered to be a pure capacitance with a possible small base conductance. The process of coupling to the probe can add inductance to the probe equivalent circuit that results in resonances that can affect the probe response. This self-resonance can often occur in the frequency band of interest.

The capacitive antennas illustrated must be electrically small, that is, the antenna dimensions must be less than one tenth wavelength at the highest frequency of interest. This restriction limits the antenna dimensions so that the maximum propagation time across the antenna is small compared with the field transient rise time, thus preserving the field waveform. Limiting the antenna dimensions limits the sensor sensitivity. The sensitivity can be increased by choosing a geometry that provides the maximum antenna capacitance for given physical dimensions.

Magnetic Field Sensors

The electrically short magnetic field sensor is a small loop. Small loops can be characterized by parameters similar to those used for probe antennas. According to Faraday's law, the voltage induced in a loop is given by:

$$V_o = \dot{B} A_e$$

$\dot{B}$ = time rate of change of the magnetic flux density

A_e = effective area of the loop.

If the flux density is uniform over the area of the loop, the effective area (A_e) is equal to the geometric area (A). The diameter of the loop must be less than one tenth wavelength at the highest frequency of interest.

A shorted loop (closed conductor) will have a short circuit current (I_s) induced which is sufficient to oppose the incident magnetic field (H), assuming no losses. This short circuit current defines the loop effective length (ℓ_e) as:

$$I_s = \ell_e H = \ell_e \frac{B}{\mu}$$

μ = permeability of the medium.

Thus a small loop will provide a short circuit current that is proportional to B, or an open circuit voltage proportional to $\dot{B}$. The effective length (ℓ_e) and the area (A_e) of a small multiturn loop are related by the loop impedance parameter (inductance) as follows:

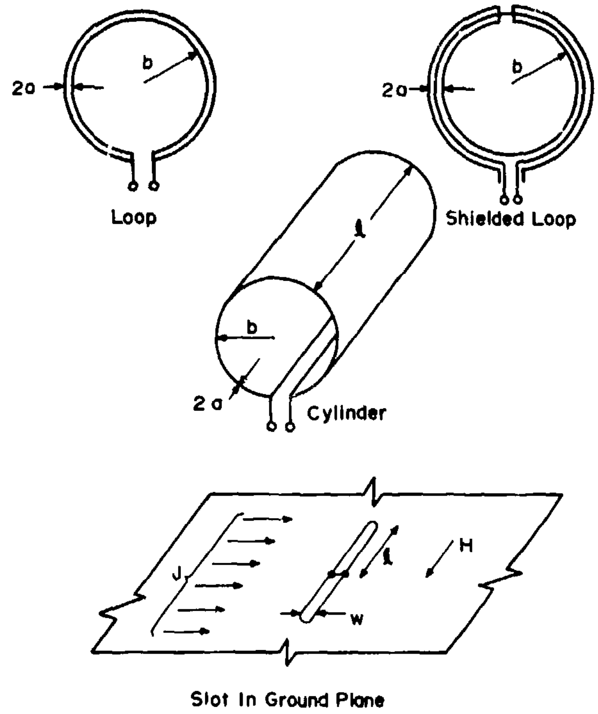
$$L = n^2 \mu \frac{A_e}{\ell_e}$$

L = loop self inductance

n = number of turns in the loop.

Illustrated are several geometries that might be used for magnetic field sensors.

An alternate technique is to measure the surface current density (J) which is related to the magnetic field, H, using a slot in a reference plane structure as illustrated above.



TYPES OF INDUCTIVE ANTENNAS

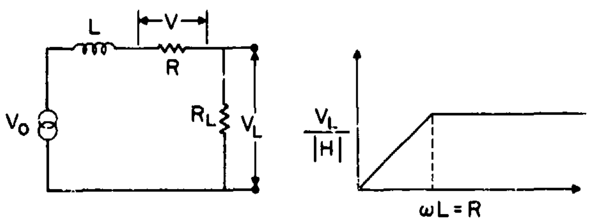
The properties of these antennas are presented in the following table.

Antenna Type	Effective Length	Effective Area	Effective Volume	Inductance	Conditions
Inductive	$\left\lvert \frac{I_o}{H} \right\rvert$	$\left\lvert \frac{V_o}{\omega B} \right\rvert$	$\left\lvert \frac{V_o I_o}{\omega \mu H^2} \right\rvert$	$\mu \frac{A_e}{\ell_e}$	
Loop	$\frac{\pi b}{\ln(\frac{b}{a})}$	πb^2	$\frac{\pi^2 b^3}{\ln(\frac{b}{a})}$	$\mu b \ln(\frac{b}{a})$	b>>a
Shielded Loop	$\frac{\pi b^2}{a[\ln(\frac{8a}{b})-2]}$	πb^2	$\frac{\pi^2 b^4}{a[\ln(\frac{8a}{b})-2]}$	$\mu a[\ln(\frac{8a}{b})-2]$	b>>a
Cylinder	ℓ	πb^2	$\ell \pi b^2$	$\mu \pi b^2/\ell$	b>>a
Slot	ℓ	$\frac{(\frac{\pi}{2})\ell^2}{\ln(\frac{8\ell}{w})-\ln 2}$	$\frac{(\frac{\pi}{2})\ell^3}{\ln(\frac{8\ell}{w})-\ln 2}$	$\frac{\mu(\frac{\pi}{2})\ell}{\ln(\frac{8\ell}{w})-\ln 2}$	ℓ> w

PROPERTIES OF SMALL MAGNETIC FIELD SENSORS

The sensitivity of a small loop can be increased by increasing the inductance L of the loop. Since the inductance is directly proportional to the number of turns and loop area, but inversely proportional to the length of the magnetic circuit (leakage inductance), it is desirable to maximize the loop area for a given length. An alternate way to increase the inductance is to increase the flux density within the loop. This can be accomplished by using a high permeability core material for the loop. High permeability cores have losses which increase with frequency, so care must be exercised or the frequency range will be restricted.

The Thevenin equivalent circuit for a loop antenna is a voltage source that is a function of the time derivative of the field and an impedance that is the loop inductance plus a loop loss resistance. This circuit is valid for a large load impedance. The impedance of this circuit is a function of frequency as is the voltage source. For frequencies where $\omega L >> R + R_L$ the impedance is inductive, giving an output voltage (V_L) that is flat with frequency. For frequencies where $\omega L << R + R_L$ the output voltage decreases with decreasing frequency. Realistically, the load impedance, R_L (resistive), is finite so that it is large compared to ωL at the lower frequencies and small compared to ωL at higher frequencies. For frequencies below that for which $\omega L = R + R_L$, the output is responsive to $\dot{B}$. Above this frequency, the output is responsive to B.



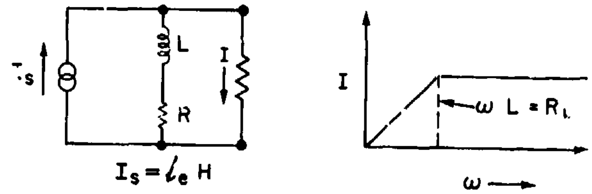
$V_o = \dot{B}\, A_e$ in Time

$V_o = \mu\, \omega\, |H|\, A_e$ in Frequency

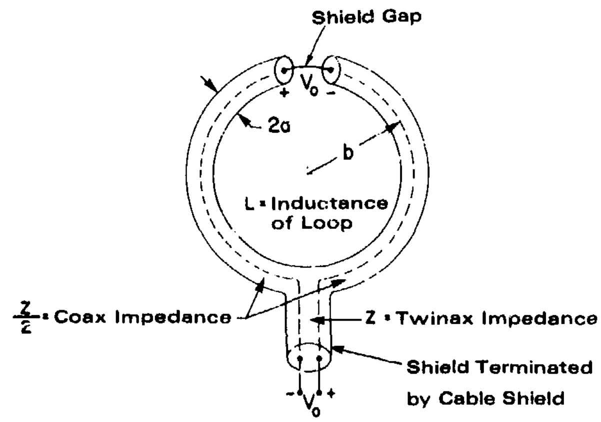
$L = \mu \dfrac{A_e}{\ell_e}$

The Norton equivalent circuit for a loop is valid when the load impedance is small (i.e., shorted terminals) compared to ωL. The output current is directly proportional to I_s which, in turn, is proportional to H. The frequency for $\omega L = R$ is the low frequency cutoff for the loop. The resistance, R, is dependent upon the loop structure, as is the loop inductance, L.

This critical frequency sets the high-frequency limit as a $\dot{B}$ sensor or the low-frequency limit as an H sensor.

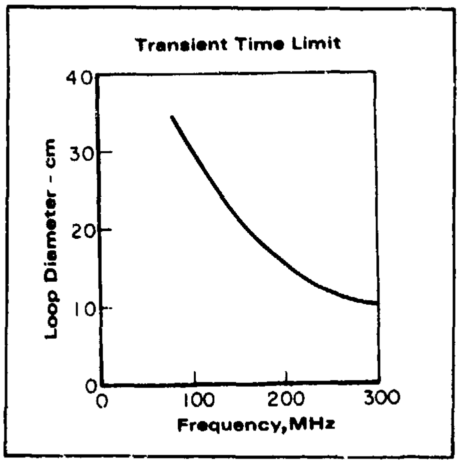


The shielded small loop magnetic field sensor is by nature a balanced structure; i.e., the loop is symmetrical in all aspects about a plane through the shield gap or load. The electrostatic shield is provided to eliminate the effects of other objects in the vicinity of the loop. The shield can be made of any highly conducting non-magnetic material. Shown is a shielded loop constructed of coaxial and twinaxial cable.

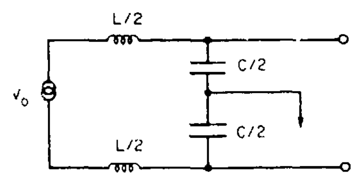


The operation of a shielded loop can be considered as the shield forming a loop that has an open-circuit voltage $V_o = \dot{B}A_e$ across the shield gap. This voltage drives the two coaxial cables in series at the shield gap, resulting in V_o at the loop output terminals.

This model of the shielded loop can be used to determine the maximum allowable size of the loop as an electrically short sensor. Essentially, the electrical length of the shield and the inner conductor must be small. The maximum loop diameter as a function of the upper frequency limit is shown here.



The presence of a shield complicates the loop equivalent circuit. The loop inductance is the inductance of a short piece of coaxial cable. This cable also has capacitance. The L and C create a resonant circuit and define the loop self-resonance. This self-resonance can be at frequencies that are within the desired loop response.



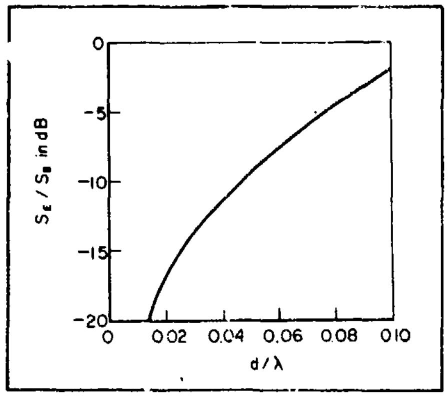
$$V_o = \mathcal{B} A_e$$

$$L = \mu b\left[\ln\left(\frac{8b}{a}\right) - 2\right] = 2\pi b \frac{Z_0}{v}$$

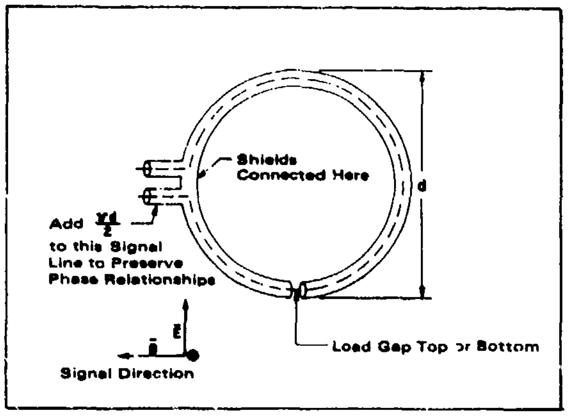
$$C \approx \frac{2\pi b}{Z_0 v} \approx \frac{3\pi b}{Z_0}$$

The loop and shielded loop are designed to respond to the magnetic field. However, loops can have an electric field response. When the electric field vector is in the plane of the loop and through the load or shield gap, the electric field response is zero. For other loop orientations, the loop will respond to both the electric field and magnetic field.

Shown is a plot of the ratio of the electric field response to the magnetic field response as a function of loop diameter. It can be seen that this ratio is quite small for small loops but can be significant for large loops.



In applications where the electric field vector cannot be made to pass through the loop load, a shielded loop can be used to minimize the electric field response as illustrated.



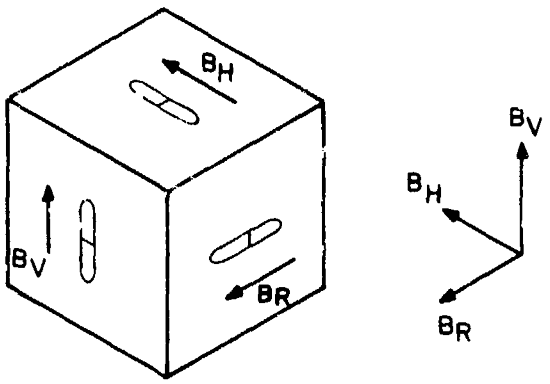
Sensor Applications

One application of electric and magnetic field sensors is determining the characteristics of the incident fields at the system under test produced by the EMP simulators.

For electromagnetic sources, both electric field and magnetic field time or frequency characteristics must be measured with care and accuracy. In addition, the electromagnetic features (i.e., spatial parameters) must be measured. Direction of propagation, spatial orientation of the electric and magnetic fields, incidence angles, phase fronts, and spacial variations in magnitude must be measured so that the testing source can be related to the EMP threat environment. Unwanted electromagnetic fields must be absent. In many instances, field components will exist in the testing source that are non-existent in the criteria.

Electromagnetic field mapping involves the use of electric and magnetic field sensors that generally have dipole response patterns. By orienting the sensor for maximum response, the electrical characteristics (time, frequency, magnitude, phase) can be measured. Rotation of the sensor for minimum response (sensor null) defines the direction of the field component. Mapping can also be accomplished using sensors that are oriented orthogonally to measure the field components in the three orthogonal directions. This method requires considerable care in interpreting the results, since sensors respond to fields in any orientation except when the field is aligned with the sensor null. For example, for a perfect sensor with zero response in the null and a single electromagnetic field component in one of the orthogonal sensor directions, misalignment of the sensor will give orthogonal components. The error in the major field component would be very small. For real sensors, the nulls are characterized by a 30 to 50 dB drop in signal amplitude. If actual orthogonal fields exist, then the sensor null response contaminates the measurement. For orthogonal fields that are within 20 dB of each other, orthogonal sensors will give reliable data.

The instrumentation that is used to map electromagnetic fields should not significantly distort the fields. This is usually accomplished by making the instrumentation as small as possible. A self-contained mapping system utilizing orthogonal sensors to measure the magnetic fields is illustrated.

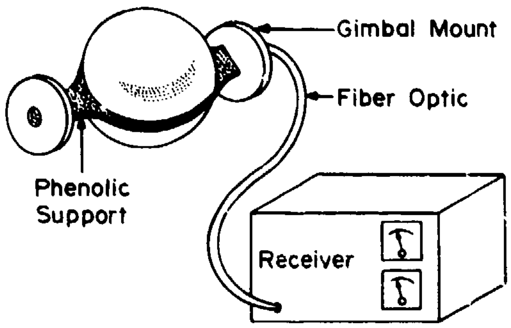


TRANSIENT MAGNETIC FIELD MAPPING SYSTEM

The recording oscilloscope and camera and battery power supplies are inside the sensor box. The sensor box provides some EM shielding for the instrumentation. Often an inner instrument shield is also utilized. If the sensor box resonant frequency is above the EM frequencies of interest, this type mapping system will not distort the measured waveforms. Calibration in a parallel plate transmission line will confirm that the measurements are not distorted. Sensor sensitivity as a function of height above ground must also be determined by evaluating the instrumentation in a known EM field.

Use of such an instrumentation box in mapping the fields in a radiating type simulator must involve consideration of ground effects on the field. Near the earth, the horizontal component of the E-field is cancelled whereas the magnetic field horizontal component is enhanced. Measurement of the magnetic field will provide the free field waveform but to obtain the actual EM field near the earth both the magnetic and electric fields must be measured or analytically determined. The following figure shows such an instrumentation box in use with a fiberglass structure to permit mapping at or above the surface of the earth.

Mapping systems can be made small enough that they do not appreciably distort the EM field measurements. A CW facility mapping system to measure electric field magnitude and relative phase and the orientation of the field as a function of height is shown in the following figure. The electric field sensor is a capacitive antenna formed by two hemispheres that contain a preamplifier and fiber optic light transmitter. The receiver is located at ground level. The small size of the sphere and the nonconducting data link and fiberglass supports do not distort the field measurements. Field orientation is determined by rotating the sphere to achieve a minimum response.



CW ELECTRIC FIELD MAPPING SYSTEM

Ideally, it is desirable to have a testing source that is not influenced by the system being tested. For limited size sources, it is often necessary to verify experimentally that this is, in fact, the case. This measurement is a part of the source definition and requires the same care and accuracy as the signal source definition measurements.

A second application of these sensors is for measurement of internal fields. These same types of sensors can be used in this application.

Voltage Probes

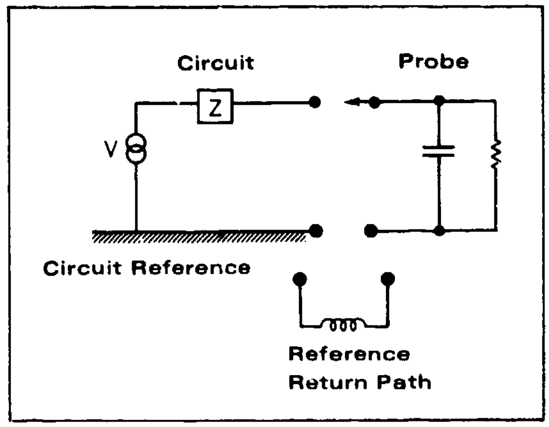
Voltage, at a point within a system, is usually measured with a voltage probe. Commercial probes are available with relatively high input impedances (high resistance, small capacitance) that can have subnanosecond response times. Often voltage probes have attenuation from input to output.

The actual response of any voltage probe is a function of the probe impedances, the equivalent impedance of the voltage source being measured, and the impedance of the circuit used to connect the circuit reference and the probe reference. The equivalent impedance of the voltage being measured is not always easy to define. Generally, voltages should be measured at low impedance locations in circuits so that the probe will not load or alter the voltage.

The reference side of the probe is usually the ground or shield side of the probe output connector. For single-ended probes, a ground terminal is available near the probe tip. A clip lead from this terminal to the circuit reference will have an impedance that is predominantly inductive.

The reference for the voltage being measured is defined by the location selected to reference the probe.

With care, low-impedance, ten's of nanoseconds rise time voltages can be measured directly. For high-impedance and/or nanosecond rise times, the high-frequency characteristics of the entire circuit, probe -- return path system must be known.



The problem of signal or circuit reference location makes it appealing to want to measure voltage directly between two points in a circuit. This is a differential voltage measurement and differential voltage probes are also commercially available.

A differential voltage probe is two single-ended probes connected to give an output that is the difference between the two voltages. The difference is determined as the difference voltage on each probe tip referenced to the probe ground or reference. Thus, the differential probe is a three-terminal network, and if the probe reference is connected to the circuit, an inductive element is again added to the measurement circuit. If a reference lead is not used as an attempt to eliminate this inductive element, then the probe reference and circuit reference will seek their own unknown return path.

Since the probe is a three-terminal network, the voltage to be measured, V, is shown as two voltage sources, V_1 and V_2, in a three-terminal network. The voltage $V = V_1 - V_2$.

The probe is designed to respond to the difference $V_1 - V_2$. To achieve this, the two single-ended voltage circuits must have identical impedances and responses. Any unbalance in the two halves of the probe will cause a common mode response; i.e., the probe output will have some response to $V_1 + V_2$. The common mode rejection ratio (CMRR) is defined as:

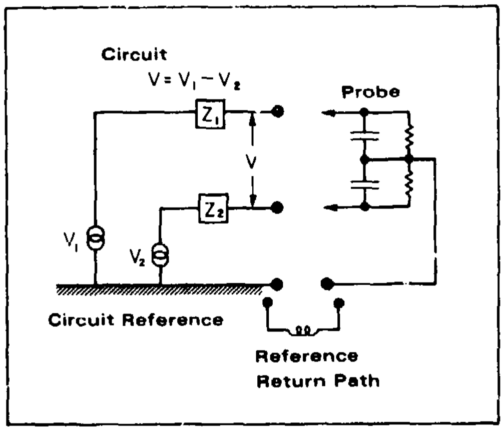
$$\text{CMRR} = 20 \log \frac{V_{diff}}{V_{com}}$$

where

$$V_{diff} = V_1 - V_2$$

$$V_{com} = V_1 + V_2$$

Common mode rejection ratios of 40 dB are attainable over wide bandwidths. CMRR's of 70 dB are attainable when carefully measured directly at the probe tip. A CMRR of 70 dB implies that the entire frequency, amplitude, phase characteristics of the probe measurement circuitry are balanced to within 0.03 percent.

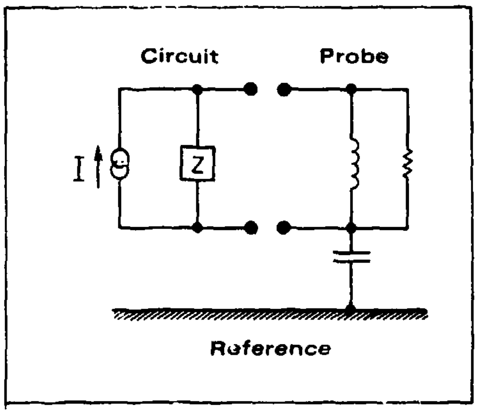


Current Probes

Current is usually measured by a current probe which is basically the secondary of a current transformer. The wire carrying the current is the transformer primary. Commercial current probes are made to clamp over a wire or bundle of wires or are made so that the wire must be threaded through the probe.

A current probe has an insertion impedance of a few μH and a few one hundreths of an ohm. This impedance is placed directly in the measurement circuit. In addition, the capacitance from the wire to current probe case adds a shunt capacitance to the measurement circuit.

The current probe, being a transformer, often contains magnetic core material to increase primary to secondary coupling. Core materials saturate and are nonlinear at high current levels. This effect of the core material can affect the time response characteristics as a function of current magnitude.



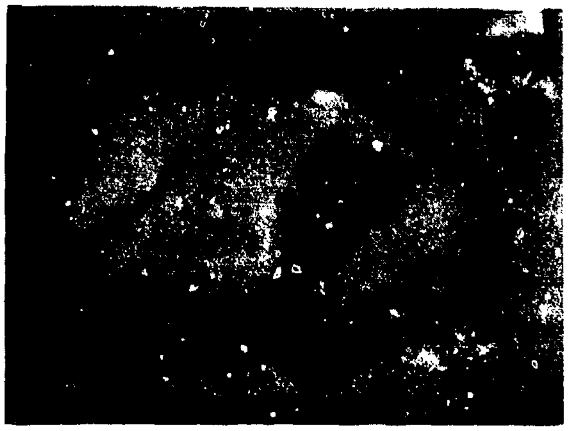
Probe Applications

Both voltage and current probes are required to define energy source characteristics. This application requires probes that can effectively measure voltage or current over the complete frequency spectrum of interest.

For cable drivers, for example, the source is a current in a cable sheath or bundle of wires. Cable driver characteristics can be determined by measuring the currents in known simple cable configurations. For voltage and current injection sources, it is necessary to know the source impedance and how this impedance interacts with the system being tested. These same simple cables can be used to determine whether the cables and cable driver interact.

Other applications of current and voltage probes are measurements of currents or voltages at the terminals or selected test points internal to the system under test. Care must be exercised in the selection of the probe to be used. The probe circuitry can detrimentally load the circuit being tested. For commercial probes, this loading is minimized and generally will not influence the measurement point. By selection of the proper point to make a measurement, it is often possible to have test points that are not sensitive to reasonable probe loading.

Bulk cable currents generally can be measured with clamp-on probes on existing system cabling. If currents are to be measured within a shielded cable, it is probably necessary to replace the system cable with a special test cable. Such a test cable is shown here with two current probes under the cable shield. Shielding integrity can be determined in the laboratory.



Signal Distribution Systems

So far the discussion has centered on measurement of the fields, voltages and currents. The sensed signals must be transmitted to a recording system. To accomplish this, some form of data transmission link must be provided. Often it is also necessary to condition the signal prior to recording. These factors are combined into a single category - signal distribution. Consequently, signal distribution systems include:

1. Cables such as coax, balanced or twinax, shielded pair.

2. Nonconducting cables such as microwave data links, dielectric waveguide, or fiber optics.

3 Amplifiers either wide bandwidths or frequency selective.

4. Attenuators and signal dividers.

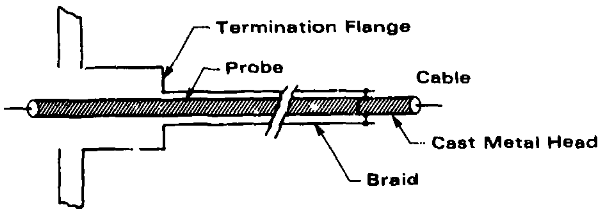
5. Signal conditioning to enhance or depress either high-frequency, short-time or low-frequency, long-time response.

These are discussed in subsequent paragraphs of this section.

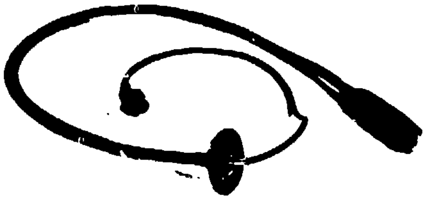
Hardwire Data Links

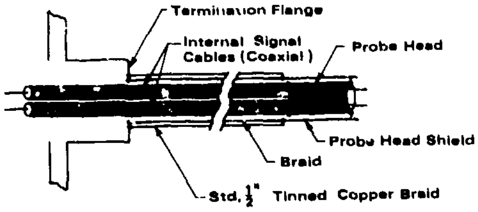
One technique for coupling the sensors and probes to the recording system is via hard wire cable. These cables can be coaxial or twinaxial and single-shielded or multiple-shielded. Most commercial probes terminate in a standard transmission line coaxial cable. The probes are designed for laboratory use and often will respond to electromagnetic energy via coupling into the probe casing. Thus it may be necessary to EMP harden the measurement system before the measurement system can be used in an EMP hardening program.

Shown is a hardened voltage probe intended for use with a recording system that is contained in a shielded container. Additional shielding in the form of braid has been added. This additional shield is connected to the probe at the sensor end.

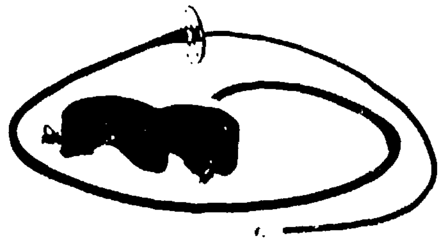


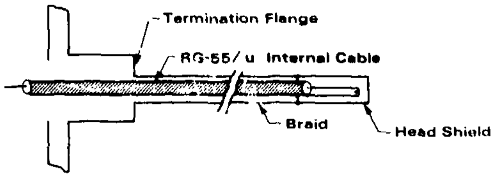
Another example, as illustrated, is the hardening of a differential voltage probe. A solid shield has been placed around the probe which contains active elements. In order to maintain a good CMRR, the cable has to be shielded with two layers of braid.





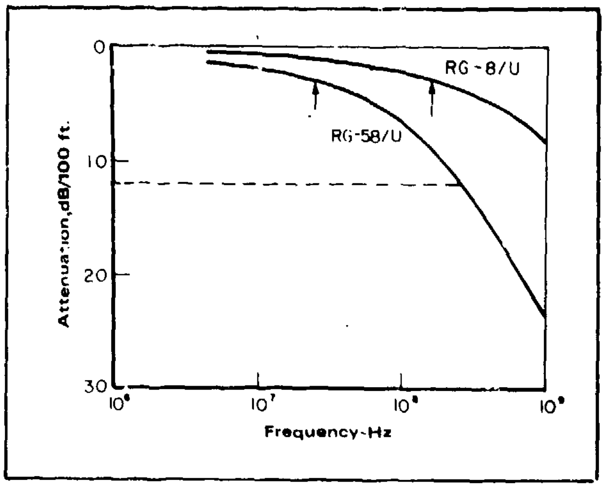
A third example is a clamp-on type current probe. This probe has a connector on the probe that is intended to be loaded with 50 ohms. Hardening of this probe involves shielding the probe to electromagnetic fields, selection of a signal cable (RG-55/U), and shielding the signal cable.





Shielding of signal cables is necessary to eliminate response to unwanted electromagnetic energies. These shields or the cable conductors themselves are sensitive to electric fields along the cable. Interference can sometimes be minimized by repositioning of cables. However, cables in electromagnetic fields alter the electromagnetic environment and can alter the system EMP interaction. Ideally, cables should be routed away from other cables and should cross other system cables orthogonally to minimize coupling to the system. Instrument cables should be routed adjacent to conducting skins or cabinets to minimize loop currents in the cables. Cables should never be allowed to dangle across open areas where they can alter the EM environment.

Hardwire cabling systems also exhibit attenuation or delay characteristics which are a function of frequency. Illustrated are the amplitude attenuation characteristics for RG/58/U and RG-8/U coaxial cables. The usable bandwidth (-3 dB points) for these cables are 25 MHz and 150 MHz, respectively. For broadband signals these characteristics must be known and accounted for in the processing of the data or by frequency compensating the cables. Synthesis codes have been developed to equalize the cable response, usually at the expense of increased attenuation as shown by the dashed line. Equalization must be accomplished for both attenuation and phase if good response of the cable is to be achieved.

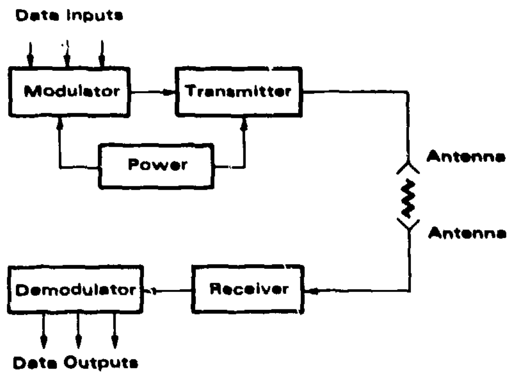


Nonconducting Data Links

Nonconducting data links can be RF telemetry, dielectric-guided RF, or optical. To be used effectively, the carrier frequency of these data links must be well above (a factor of ten or so) the EMP spectrum. Segmented and loaded (lossy) transmission lines have limited bandwidth and are not commonly used.

RF Telemetry

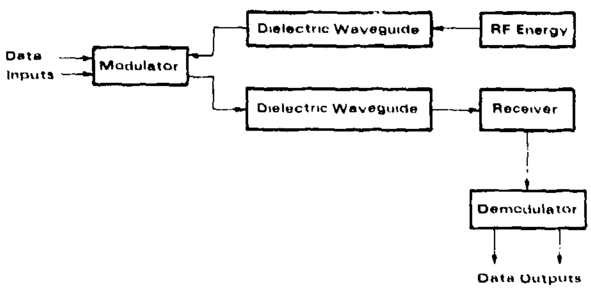
An RF data link can take many forms. Commercial microwave data links are available that can be altered for broadband transient data transmission. Essentially, such a data link consists of an energy source, modulation from simple single-channel AM to multiple-channel FM-FM, directional antennas (usually a horn), a receiver, and a demodulator. Power in the form of batteries must be supplied at the transmission end of the link. Thus, the physical size of the instrumentation that must be added to the probes can overcome any advantage of the nonconducting data link.



This form of data transmission requires radiation of an RF signal. If the RF signal is within the EMP spectrum, the radiated (simulator) field will interfere with, and may even mask, the telemetered data. Choosing a telemetry frequency far above (ten times the highest frequency in the EMP spectrum), low pass filtering can be employed to eliminate this interference problem.

The radiated RF telemetry signal can also interfere with the measurement of the radiated EMP fields. For example, if the EMP field sensors have a very broad bandwidth, they may respond to the telemetry signal resulting in distortion of the measured EMP fields.

The physical size of the probe end of the data link can be reduced if power is supplied from the receiver end of the link. This may be accomplished by providing the RF power from the receiver end via a second channel (as illustrated) or two way propagation on a signal channel.



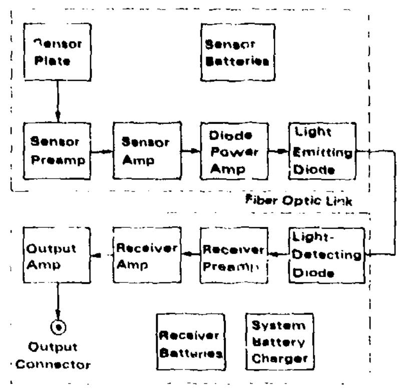
Dielectric Waveguide Transmission

The use of dielectric waveguide, illustrated above, eliminates the need for radiating antennas. Energy propagates both within and on the rod surface in a transverse magnetic mode and may extend 1 to 6 inches beyond the surface depending on the frequency and dielectric material. Thus, some protection is required to keep conductors away from the rod. Propagation losses less than 0.1 dB/ft. have been measured at 12, 24 and 48 GHz. Dirt on the rod surface can increase this loss.

Dielectric rod transmission systems have been successfully used at Ku (12 to 18 GHz) and X (8 to 12 GHz) band frequencies with rod sections in excess of 100 feet. Bending of the rod on a 10 foot radius does not noticeably affect system performance.

Optical Transmission

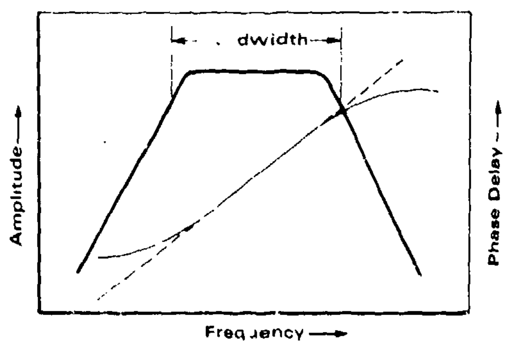
Data can also be transmitted by modulating light. The light-emitting diode offers an easy method to modulate light, its output being directly proportional to the current through it. The transmission link can be lens and mirrors; however, mechanical alignment limits their field application. Light can easily be transmitted using flexible fiber optics. The size of an optical data transmitter can be quite small. Optical links with bandwidths up to 150 MHz are commercially available with relatively low loss.



Signal Conditioning

Amplifiers

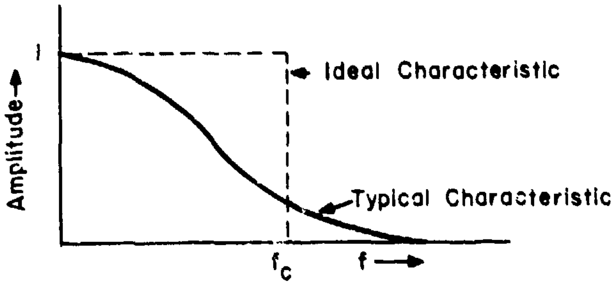
Amplifiers for pulse or transient application have a constant gain and linear phase response over a broad range of frequencies defined as the amplifier bandwidth. For linear phase response, the phase delay increases linearly with frequency. This is equivalent to a constant time delay for all frequencies which results in no dispersion or phase distortion of a time waveform. Outside the amplifier bandwidth, amplitude will vary with frequency and phase will deviate from linear phase. This causes pulse distortion in an amplifier.



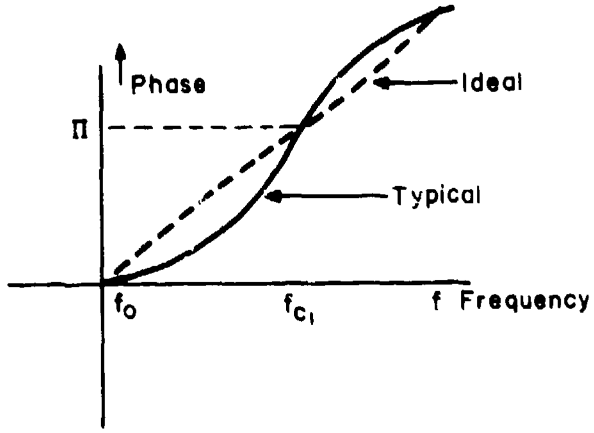
Both the low-frequency cutoff and high-frequency cutoff characteristics are important. These cutoff frequencies are determined by the RC time constants associated with the amplifier circuit.

The high-frequency response of the amplifier limits the rise time of the output waveform. The rise of the output waveform is exponential. If the time constant ($\tau 2$) associated with the high-frequency cutoff is less than five (5) times the rise time (t_1) of the input pulse, the error introduced will be less than one (1) percent. The high-frequency response also controls the decay of the pulse which will also be exponential. Inadequate high-frequency response will result in a reduction in the output pulse amplitude.

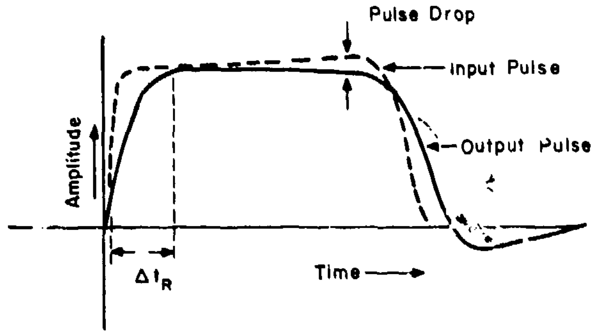
The low-frequency response of the amplifier results in a sag in the top of the pulse (output voltage dropping due to changing of the capacitance associated with the output coupling circuit). This effect is seen primarily for long pulses, (i.e.. where the pulse duration is long compared to the RC time constant of the output circuit). The low-frequency response also results in overshoot at the end of the pulse due to this charging of the coupling capacitance.



In addition to the amplitude response, distortion of the output pulse can also result from non-linear phase delay of the amplifier. The phase delay distortion for the first-distortion terms in the Fourier representation are shown in the figure. The dotted line shows the ideal characteristic (linear phase delay with frequency) and the solid line the sinusoidal deviation from the ideal.

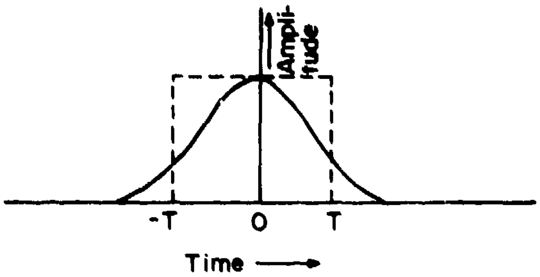


The effects of the high and low-frequency cutoff characteristics of an amplifier are depicted qualitatively in the following figure. The figure indicates the pulse droop or sag due to the amplifiers low-frequency cutoff and the degraded rise time of the output pulse due to the amplifiers high-frequency cutoff.

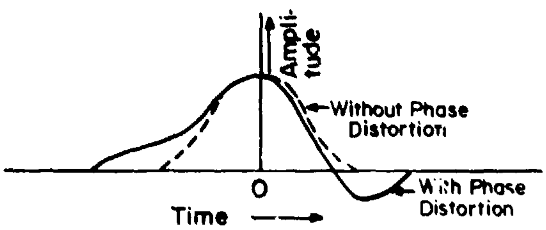


The low-frequency and high-frequency cutoffs do not amply define the response of a broadband amplifier, since the relative amplitude and phase of signal spectral components outside these limits affect the amplifier response. Shown in the figure is the ideal amplifier amplitude response characteristics (dotted line) and the typical pulse response (solid line).

The effect on the output pulse shape for a square wave input pulse is shown in the following figure. Figure (a) shows the distortion due to the amplitude response only, and Figure (b) includes the effects of phase distortion.



A.) AMPLITUDE RESPONSE



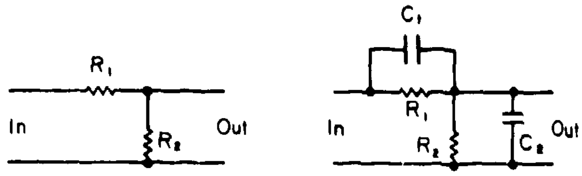
B.) EFFECTS OF PHASE RESPONSE

Amplification of CW signals can take a variety of forms. These can include simple straightforward amplification with bandwidth-limiting filters to reduce noise. Frequency translation can be used but must be done coherently to preserve phase. Detection of amplitude and phase of the signal require coherent detection that can be accomplished with synchronous detectors in phase quadrature.

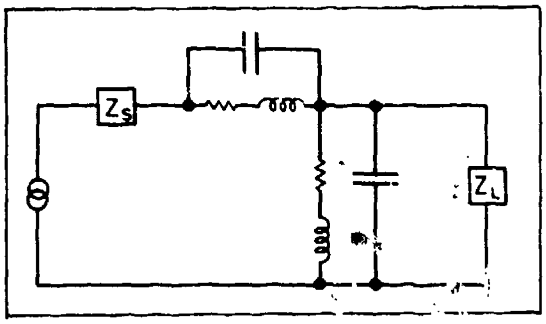
Attenuators

The concept of attenuation is a simple one. If two resistors are connected, as shown on the left, then the output voltage is attenuated by $R_2/R_1 + R_2$ from the input voltage. If the resistance values are not too large and frequency is low, this form of attenuation is valid.

However, as frequency is increased, the stray capacitance associated with the output (C_2) will alter the attenuation ratio. Its effect can be compensated by the addition of C_1 in the compensated attenuator shown. Compensation required to maintain the desired attenuation ratio is given by $R_1C_1 = R_2C_2$.

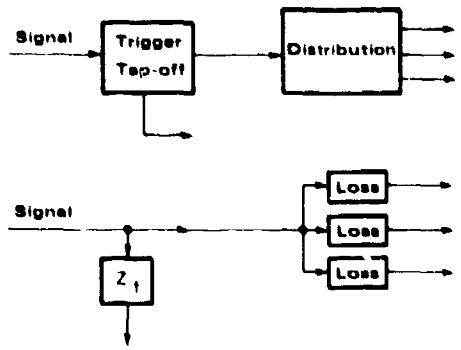


For still higher frequencies, all lumped components have R,C, and L. In addition, the voltage source will have a source impedance and the load a load impedance. The effects of all these impedances must be considered when designing high frequency broadband attenuators. Any attenuator, no matter how complicated its compensation may be, has some useful limited frequency range or bandwidth.



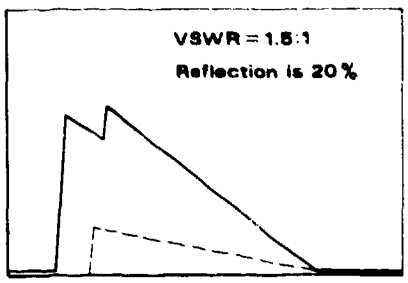
Signal Dividers

Often, between the sensor or probe and data recorder, the signal is distributed either to multiple recorders or to perform other functions, such as triggering an oscilloscope. These functions must be accomplished without distorting the signal channel. Signal energy for triggering can be obtained with a high impedance tap off whose impedance is much larger than the signal channel impedance. Signal distribution is normally accomplished with a signal divider that has an input impedance that matches the signal line impedance and generally will have individual output impedances matching the output line impedances. Passive circuits that have constant input and output impedances have isolation between input and output in the form of loss or signal attenuation.



Impedance mismatches cause reflections in the signal channel. For CW, the reflections result in a voltage standing-wave ratio (VSWR). A VSWR of 1.5, which is considered adequate for some applications, corresponds to a voltage reflection of 20 percent. As indicated here, a 20 percent reflection can result in considerable distortion of a transient.

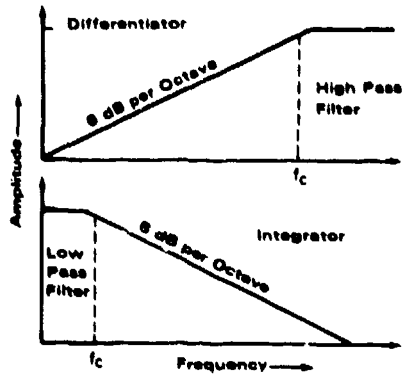
One technique that is useful to define and locate reflections in a signal channel is time-domain reflectometry. Commercial TDR's are readily available with subnanosecond accuracies.



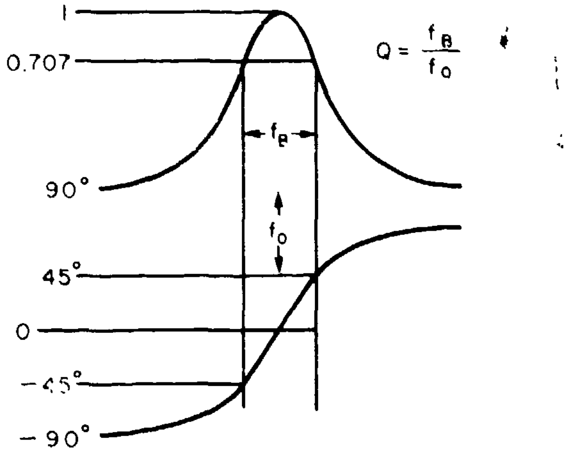
Differentiators and Integrators

Sometimes it is desirable to distort the signal spectrum intentionally. Differentiation of the signal will enhance the high-frequency response or result in a signal that is the time derivative. Differentiation can improve time definition. A differentiator has an amplitude response that is directly proportional to frequency. A single-stage high-pass filter shown in the upper part of the figure has an amplitude response variation of 6 dB per octave below the cutoff frequency, f_c. Thus, for frequencies below f_c, the output is the differential of the input. Similarly, a low-pass filter with 6 dB per octave roll off will integrate all frequencies above f_c.

For broadband applications, f_c must be well outside the signal spectrum. In addition, the effects of source and load impedance and distributed impedances will limit the useful frequency range of a true integrating or differentiating circuit.



Other forms of special filters may be used either to enhance a frequency or frequency band or to attenuate a discrete frequency or frequency band (notch filter). Such techniques should be used with caution since phase as well as amplitude are altered over a frequency range that is the filter bandwidth.



Signal Display and Recording

Recording of transients in the time domain is normally done using an oscilloscope and film. A single recording or multiple recordings to extend dynamic range may be used. Transient recording can be digital or a simple peak magnitude recording.

CW is usually recorded on chart recorders that can be simple time-amplitude recordings of x-y plots. Often the recording of a meter deflection will suffice.

Oscilloscope/Camera Recording

In spite of recent developments in alternative recording techniques, the best instrument for recording a transient is still the oscilloscope-camera combination. Oscilloscopes are normally specified as to bandwidth, deflection sensitivity, sweep time, and writing rate. These parameters define an oscilloscope but do not define its capability to display transient data.

For the display of transient data, it is convenient to define the oscilloscope capabilities in terms of the display spot size. The time response of the oscilloscope depends upon sensitivity in volts per spot width and the slewing rate in deflection per unit time. Slewing rate depends on oscilloscope writing rate and upon the volts per second capability of the amplifiers. Resolution is defined in the number of tracewidths that can be displayed. The tracewidth is the minimum width of the trace of an oscilloscope.

OSCILLOSCOPES

SPECIFICATIONS

- BANDWIDTH
- DEFLECTION SENSITIVITY
- SWEEP SPEEDS
- WRITING RATE

TRANSIENT RECORDING

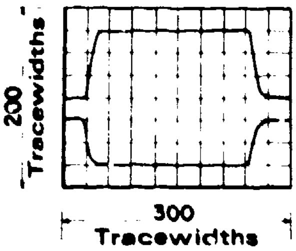
- SPOT SIZE
- SENSITIVITY
- SLEWING RATE
- RESOLUTION

Typically, the vertical deflection is limited to about 200 tracewidths. This number of tracewidths limits the dynamic range to a maximum of 46 dB, providing that a change of one tracewidth is observable in data reduction.

The recorded bandwidth is determined by the number of tracewidths along the time axis, typically 300, and the sweep duration. A theoretical limit of definition five times this minimum sampling rate is desirable; i.e., 10 tracewidths per cycle. Thus, for an oscilloscope with 300 tracewidths per sweep, a sweep speed can be selected to display 30 cycles of the highest frequency to be displayed. This length of sweep will display one cycle of a frequency that is 1/30 of the highest frequency, which is the lowest measurable frequency that is recorded. Thus, 300 tracewidths limit the recorded bandwidth to a maximum of 30 to 1.

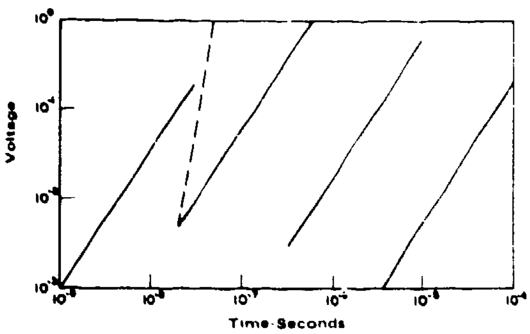
A 150-MHz oscilloscope set to record 150 MHz will not display frequencies below about 5 MHz. If the sweep speed is set to record 10 kHz, then the maximum observable frequency is less than 1 MHz, even though the oscilloscope is capable of responding to 150 MHz.

Typical Oscilloscope Display

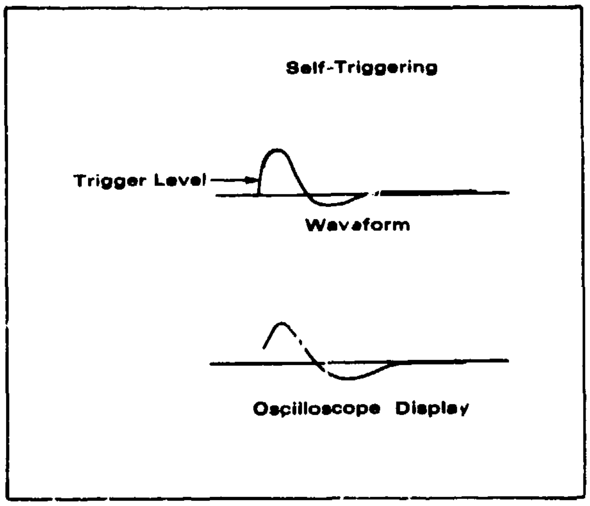


A single oscilloscope recording is very limited in its amplitude and time dynamic range. Usually, to record transients associated with EMP hardness testing, more than one recording of each waveform is required. The recording dynamic range sets a time-amplitude window that can be recorded. Four such typical windows are shown here as the solid lines. The upper end of each line represents the full-scale deflection voltage and the sweep length. The lower end of each line is the voltage and time resolution for the particular oscilloscope settings.

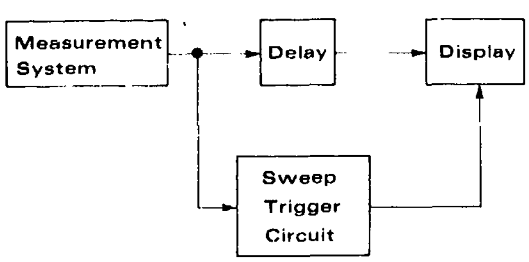
The dashed line represents a rapid sweep recording that has been initiated at a later time by a delayed trigger. This technique permits good time definition at latter times in the transient waveform.



To display a waveform on an oscilloscope requires that the oscilloscope sweep and the waveform be aligned in time by triggering the sweep at a known time with respect to the waveform. Triggering can be derived from the waveform, thus starting the sweep after the waveform onset. It is important to realize that, if the oscilloscope sweep is nonlinear, it is probably nonlinear at the sweep onset. Thus, with self-triggering, the initial portion of the waveform is lost and may possibly be distorted.

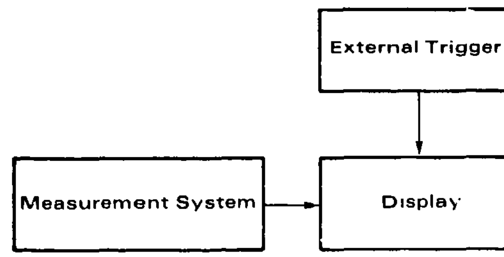


This problem can be overcome by adding delay to the signal channel that is not in the trigger circuit. Many oscilloscopes have internal delay for this purpose. However, the signal distortion characteristics of broadband delay lines must be included in the overall measurement system response. Delay of a few ten's of nanoseconds can be obtained using good coaxial cable.



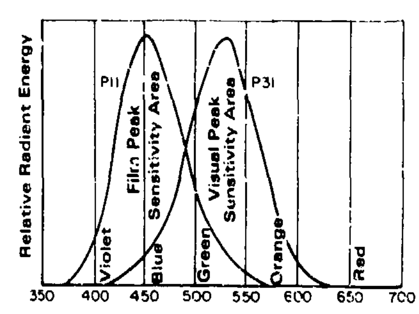
An alternate method of aligning the waveform and oscilloscope trace is to provide an independent trigger to initiate the sweep prior to the waveform. For repetitive pulse applications, this trigger source can be the waveform itself with delay in the trigger channel that is slightly less than the period between pulses.

For transient recording, the trigger can be derived from the fields that impinge on the system or from the sensed simulator field. Either of these methods can contaminate the measurement itself. Extreme care must be used to ensure that there are no unwanted EMF penetrations into the measurement system.



The recording of a waveform or transient on film is dependent upon many factors. Trace recording is ideally line photography, so that overexposure of the film seems to be desirable. However, the oscilloscope trace has a Gaussian distribution that will cause the trace to broaden as trace intensity is increased. The effects of high intensity are not always obvious to the eye since the eye responds logarithmically over a million-to-one range, while film responds linearly over a 100-to-one range.

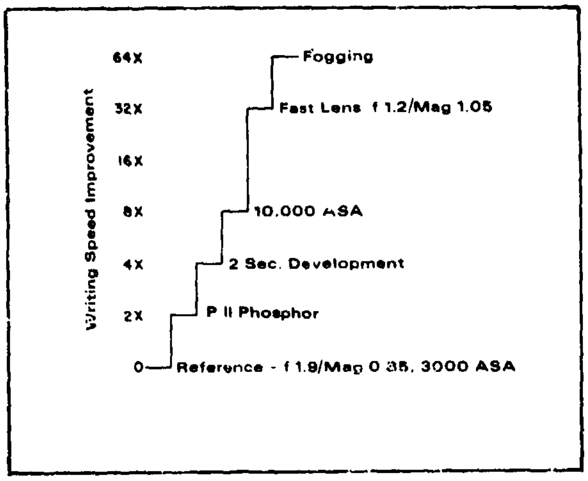
Most oscilloscope cameras will have at least an f 1.9 lens with a slight image reduction (magnification = 0.85). The most common phosphor for the oscilloscope is P-31, which is optimized for visual use. Other phosphors, P-11 for example, are better matched to films that are sensitive generally to blues.



The film writing speed is a function of many factors. These factors include:

- Oscilloscope tube phosphor
- Development time
- Film speed
- Lens f number (speed)
- Film pre-processing

The increase in writing speed that can be obtained by proper choice of these factors is shown in the figure.



The exact form of film and camera that is used will depend upon the data interpretation process that is used with the test program. Polaroid cameras and film give an instant permanent record. The recording can be viewed to determine if it is readable and if additional records are required. This feature eliminates the recording of useless data. The disadvantage of Polaroid film is that it is difficult to use in machine processing including viewers and only a single copy (print) is available. Polaroid does make a negative film intended for use in making slides; however, the speed of this film limits its usefulness in transient recording.

The Polaroid camera is essentially a framing camera where the film can be positioned for multiple exposures if desired. Framing cameras exist for standard films, usually 4 x 5 inches.

For measurements that are repeated many times, a stepping camera is useful. Stepping cameras are usually 35 mm and automatically stepped one frame at a time. The film is continuous.

Continuous motion cameras can be used similar to a stepping camera, providing the film motion distortion during the oscilloscope sweep is not objectionable.

The improvement in sensitivity that is possible by using CW and narrowband filters can be utilized to display repetitive pulse measurements. Current state-of-the-art sampling techniques have been utilized with standard oscilloscopes to display sine waves about 10 GHz and repetitive pulse rise times in the ten's of picoseconds (10^{-12} seconds).

Sampling is a process wherein a narrow-time gate is used to measure and store the input signal that occurs at the time of the sample. By moving the sample gate along the waveform on subsequent repeated waveforms, the entire waveform is sampled, stored, and displayed. Sampling at each point on the waveform is normally a single sample; however, multiple samples are possible so that averaging can be used to improve signal-to-noise ratios.

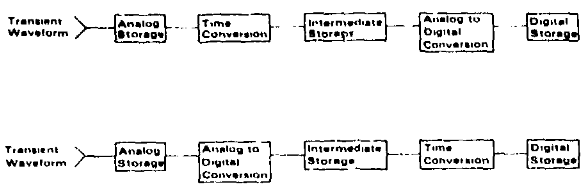
Analog/Digital Recording

Analog recording of a transient can be accomplished on tape recorders or on magnetic discs. Bandwidths of ten's of megahertz are feasible. However, tape or magnetic recorders have poor signal-to-noise ratios, limiting their dynamic range to about 30 dB. Linearity through playback can be adequate with proper calibration. The overload characteristics of tape usually demonstrate poor recovery.

Various methods are available to record transient waveforms in digital form. The building blocks are available in various forms so that particular systems can be assembled for specific applications.

Generally, digital recording involves analog storage to permit waveform sampling, analog-to-digital conversion of amplitude, conversion of the time scale, intermediate or temporary storage between the time and amplitude conversion processes, and finally, storage of the digital information. The most common version of digital recording is shown in the upper diagram. The analog signal is stored in some manner so that time samples can be determined (usually at a nonreal time) and then each sample is digitized and stored. The time conversion permits the analog-to-digital conversion to be accomplished at a lower bit rate.

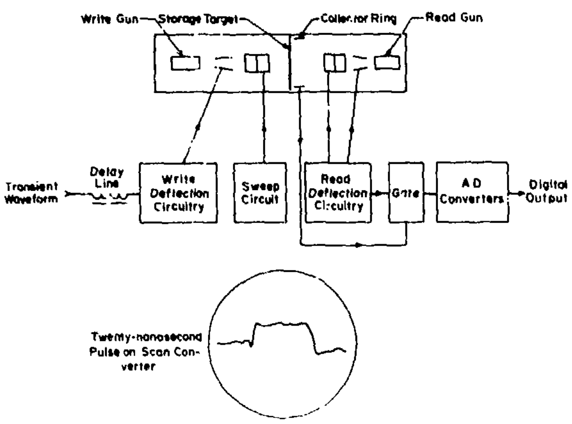
An alternate approach is shown in the lower diagram where the analog-to-digital conversion is accomplished first. Time conversion is then accomplished on the digital data. This technique is generally limited in bandwidth by the analog-to-digital sampling rates, or more correctly, bit rates.



One digital recording technique -- <u>the scan converter</u> -- has been widely <u>used to record transients</u>. This scan converter utilizes an oscilloscope with a precharged semiconducting target that retains the oscilloscope trace as a charge pattern resulting from secondary emission initiated by the oscilloscope electron gun. A second electron gun is then used to scan the stored pattern (time conversion) at a slower rate for conversion to digital form.

The waveform shown is the analog output from a scan converter read gun prior to analog-to-digital conversion. The waveform is 20 nsec wide but is displayed over a longer time.

THE SCAN CONVERTER



One technique that has been used in EMP measurements is to make many measurements of a single feature of the EMP transient. The peak amplitude of a waveform can be recorded on magnetic film by recording a carrier frequency on the film and using the waveform to be measured to erase or partially erase the film.

Energy thresholds can be recorded using such simple devices as fusible links.

<u>CW Recording</u>

Recording of amplitude, phase, and frequency for CW measurements can be accomplished in a variety of ways since the recording bandwidth for a single data point is very small. Oscilloscopes can be used to display CW data but generally are not used due to the film processing and reading.

The most common form of CW recorder is a pen-chart recorder. These recorders are available in many forms with the writing mechanism being ink, electric, thermal, optical, or impact. Recording bandwidths in the ten's of hertz are easily attainable and adequate. Chart recorders can be single-channel or multiple-channel.

X-Y recorders can be used to display CW data in a form that is more acceptable for human interpretation. An X-Y recorder is essentially a special-purpose chart recorder.

The normally slow data rate for CW data allows the use of simple analog-to-digital converters. The incremental digital tape recorder is well suited to stepped low-data rate data recording. The recording of CW data in digital form is useful if further machine processing warrants its use, since machine processing will be required to provide a visual display that chart recording will give directly.

<u>Data Processing</u>

Raw recorded data is useless unless it can be displayed in a meaningful way. Good data processing unifies a set of disjoint raw data points into a picture of how the system responds to electrical transients. There are many reasons for processing raw data.

Among these reasons are:

- To expand the dynamic range of a single recording by the proper recombination of multiple recordings.
- To indicate trends or determine specific responses
- To alter the data format, such as conversion from the time domain to the frequency domain, etc.

- To average or summarize the results from a number of discrete tests to minimize measurement errors
- To compare data
- To average out noise.

There are a wide variety of techniques or machines to perform these data reduction or processing functions. In some instances, where only a limited amount of data must be processed or where only a "quick look" is desired, the processing may be accomplished manually. Manual processing may include manual digitization of oscillograms, etc. Where large volumes of data must be processed or where a great deal of manipulation of the data is required, manual processing is impractical and machine processing must be employed. A few typical categories of machines for this purpose are indicated.

Processing Machines

- Analog-to-digital scalers
- IBM card systems
- Plotters
- Special-purpose computers
- General-purpose computers

Where machine processing is required, the necessary programs must be available. Some of the more commonly used are indicated. Special programs may have to be developed depending on the type of data, data format, or information output desired from the data. A great many special programs exist but are not listed since they generally are not applicable.

Mathematical Tools

- **Time Domain Conversion**
- **Statistical**
- **Circuit Analysis**
- **Special-Purpose**

Calibration

Proper calibration necessitates a complete understanding of the response characteristics of the entire measurement system. For transient recordings, the transient or pulse response of the system must be measured and documented; and for CW measurements, the frequency vs. amplitude and sometimes vs. phase must be measured and documented. These calibrations must be periodically verified.

Calibration signals should be included with the data to be certain that recorded amplitudes, times, and frequencies are correct. Calibrations should be an integral part of the entire data measurement procedure including the data reduction system. Recorded data is useless unless the relationship between the signal being sensed and the tabulated or displayed data is known.

Calibrations determine overall response characteristics, relate inputs to outputs, and provide scale factors for the data. In practice, measurement systems calibrations are accomplished in three steps: in-the-field measurement system calibration, data record calibrations and data processing calibration.

Calibration, ideally, should only have to be performed at the onset of the test program. This would be acceptable if the simulator output, and the measurement system response did not change with time. In some cases, changes have been seen to occur from shot to shot. Therefore, periodic checks are required. The frequency at which the calibration must be repeated is a function of the performance of the energy source and the measurement system. This repeatability of data must be determined prior to the actual test program.

Factors which cause instrumentation system calibration to change include:

- Carelessness resulting in damage to the instruments or telemetry links.
- Misalignment or improper connection of telemetry links
- Temperature effects
- Aging of components or batteries used in remote units
- Rerouting of cabling and changing length of telemetry cables
- Replacement of signal wires or components.

Many of these factors are long-term maintenance problems. Others, such as battery charge conditions, may change over short periods on the order of 2 to 4 hours. Again, depending on the known performance characteristics and condition of the system, the frequency of calibration checks can be established.

Measurement System Calibration

The measurement of the in-the-field system performance involves exposing the measurement system to a known signal and recording the resulting output signal. For probes, known voltages and currents can be directly coupled to the probes. For a field sensor measurement system, the entire system must be exposed to a known electromagnetic field. This often is done using a transmission line simulator as discussed earlier. If a probe measurement system is to be used in the presence of large simulated EMP electromagnetic fields, then that system response to the incident fields must be known.

The impulse response or CW amplitude and phase response can be used to define system measurement characteristics since, for linear systems, the impulse response is the inverse Fourier transform of the CW amplitude and phase response. If the CW response is to be used to unfold time history recordings, response data is required to define amplitudes that are down at least 40 dB from the maximum or mid-band response. Phase must be defined at each frequency. Phase measurements must be made at a minimum of five frequencies per cycle of system phase response ($0 < \phi < \Pi$ radians). Both the phase and amplitude must be measured at all frequencies.

The accurate measurement of phase using an oscilloscope to record the data is difficult. Determination of the zero crossing or specified amplitude level on successive cycles from the oscillograms can easily result in errors due to manual data reduction. Also, low-level CW signal response cannot be used to determine measurement system overload response which is a function of signal amplitude and is a nonlinear process.

Pulse or transient measurements can be used to determine system response. The step function response and/or impulse response can be measured, providing sufficient care and accuracy are used to make the measurements. A step function input must rise in a time that is at least an order of magnitude faster than the response rise time of the measurement system, and must remain at constant amplitude for a time that is long compared to the longest observable time limit of the system determined from the system low-frequency response. An impulse must have a total duration that does not exceed 1/10 the measurement time response or rise time. These requirements on the waveform used to determine system response means that the waveform spectrum exceeds the measurement system bandwidth. Observation of these input waveforms requires the use of even wider bandwidth oscilloscopes. For repetitive pulses, wider bandwidths can usually be recorded using such techniques as a sampling oscilloscope.

The system transient response measurement demands that multiple records of the measurement system output be made with a wide range of amplitudes and time bases so that the output time history can be accurately defined. Definition of the measurement system pulse response means that the system time response is defined for all times, both fast rise times and long decay times. The response must be measured at a sufficient number of time points so that straight lines between the points still define the waveform. Often, over 1000 points are required to define each waveform accurately.

The time response of a system can be measured using any other input waveshape, provided the waveshape is accurately known and its spectrum is well known over a bandwidth that exceeds the measurement system bandwidth. The response of a system can never be defined by a testing spectrum that only partially covers the system bandwidth.

Repetitive pulses with known spectral notches or amplitude zeros should not be used to determine system response because of the obvious problems associated with division by a vanishingly small number.

Data Recording System Calibration

An in-the-field measurement system designed to record waveforms whose response is well known is useless if the system produces unmarked, unlabeled photographs or charts. Data calibrations must accompany each record so that each point of the record can be assigned a value that can be related to the sensed signal using the known measurement system response. Most receiving and recording systems will have nonlinearities which must be calibrated even though the system may be reasonably linear in the region in which it is intended to be used.

Waveform calibrations required include a standard time waveform. A CW signal is often used for a time reference. CW frequencies that are above the normal system bandwidth must be used to adequately quantify short times. Frequency can be accurately measured to determine time.

Amplitude calibrations can take the form of volts-per-unit deflection or can be applied at the measurement system sensor or probe to give directly the sensed parameter-per-unit deflection. If a fast-rise, flat-topped pulse is used for waveform calibration, care must be exercised to select an amplitude point on the calibration waveform where the system loss is well known to calibrate the data display in terms of volts/division or other parameters since pulse distortion may be present. A CW signal within the system response bandwidth can be used to alleviate this problem. However, a single frequency does not verify that the waveform system response is unaltered. It seems desirable to use some combination of pulse and CW for calibrations.

Regardless of the type of amplitude calibration that is used, more than one amplitude is required to check for nonlinearities. Just as important, it is necessary to know zero level for all readings since amplitude is a linear measurement from some reference (zero) to the point on the waveform.

Waveforms that have both rapid and long-time variations are often recorded on multiple records. The assembling of these records into a single waveform record requires that overlapping features of the individual records be matched -- usually in time. Waveform features that are well defined can be used but often do not exist where wanted. Thus, some common time reference is desirable for multiple recordings. Trace intensity modulation can be used for this function. A calibrated waveform also requires labels to define the measurement parameters that are not a part of the measurement system. These housekeeping parameters are test point location, date of measurement and other parameters that relate the data to the EMP hardness test plan. Often included are measurement system parameters, such as control settings, that can be used to verify the calibrations or sometimes in place of waveform calibrations. The redundancy in housekeeping records can resolve human errors in the data.

Measurement systems that record data in the frequency domain must also provide calibrated records. Both the amplitude and phase result in some form of recording deflection. Each recording requires calibration. For amplitude-phase calibrations, it is often desirable to inject the calibration signal at the sensor or probe so that both amplitude and phase can be related directly to the measured signal. If calibrations are injected at any other point in the measurement system, then the system response calibrations must also include the calibration system response.

Amplitude calibration is usually performed using a signal with a fixed known phase. Phase calibration is performed with a known constant amplitude signal. The system phase response may be a function of amplitude, thus requiring additional data calibrations with several signals of different amplitudes.

The calibration signal must be precisely located in frequency relative to the frequency characteristics of the measurement system. For relatively broadband systems, this is simple. However, if the system has a very narrow bandwidth, it may be necessary to provide within the system some method whereby the calibration signal and system tuning frequency can be aligned by zero-beating, for example. In all cases, the data records require an indication of system frequency and calibration signal frequency.

For CW data recordings as with all data recordings, zeros of amplitude and phase must be included in the data calibrations. Since normally more than one recorder is required to record amplitude, phase, and frequency, some common reference to all recordings is required. For single-drive, multiple-recording channel recorders, the known relative positions of the various recording pens suffice. For multiple recorders, a common reference signal and "side marking" pens are commonly used. As with any recording system, housekeeping functions must be recorded to define the measurement and the measurement system.

Data Processing System Calibration

Knowing the in-the-field measurement system response characteristics and having well marked and calibrated data records are necessary but are not sufficient to a good test program. The response characteristics of the data processing system must be known since processing translates the recorded data into a meaningful format.

Even the simple processing that is involved in hand plotting of data will alter the dynamic range and reading sensitivity.

Processing calibrations must be made for time base translation, time-to-frequency translation, for frequency translation, frequency-to-time translation, and for amplitude translation. Linearity, dynamic range, and sensitivity must be determined.

Data processing is an information filter and, as such, can only lose information. Information is lost in processing errors. Processing errors can come from machine and human error or from incorrect or inadequate mathematical processes. The processing of representative sample data or sample problems will verify calibration of the processing tools.

Test Instrumentation Setup

Many instrumentation concepts have been presented for use in an EMP test program. The implementation of these techniques, some of the precautions that must be taken when implementing these concepts, and approaches to insure the data obtained is reliable and valid are discussed in this section.

Testing Requirements

System hardness testing requires the use of simulators to create EM fields. The measurement of these fields over the test area (mapping) must be done with care. The test instrumentation must be checked for noise immunity. Further, the instrumentation must not influence the response of the system under test. It is not always easy to differentiate between valid data and noise.

The system that is under test must be exercised for all operational modes and configurations. The operational performance of the system must be ascertained before, during and after the test. Test configurations include system orientation in the EMP field, and determination of points of entry into the system.

In the early stages of a test program, the experimental determination of the orientation and operational mode that gives the maximum EMP interaction with the system is required.

During the test planning stage, the various portions of the system must be reviewed to determine the electromagnetic response of the system. Each portion of the system will have a maximum response for some specific orientation of the applied electromagnetic field. This maximum will not necessarily occur on all portions of the system at one particular field orientation. Thus, it is often necessary to determine experimentally if a particular field orientation is the worst-case orientation. If a worst-case orientation can be found, then all subsequent testing can be accomplished at that orientation. Predicted worst-case orientations should always be experimentally verified.

The operational modes of the system can also effect its response. Therefore, verification of predicted worst-case operational modes is required. Generally, to determine the worst-case operational mode involves functional testing or damage testing.

Critical point testing is the experimental determination or verification of the critical points in the system. Two types of tests are used. One is an internal mapping of signals at the various entry points to determine those areas within the system that give maximum response to EMP fields. Initially, such mapping can be accomplished with quick looks at the data to rank-order the observations. This rank-ordering can often be done with a visual observation.

The other procedure for locating critical points is to test portions of the system in the laboratory and observe damage or upset. The mating of these two experimental approaches defines the system critical points.

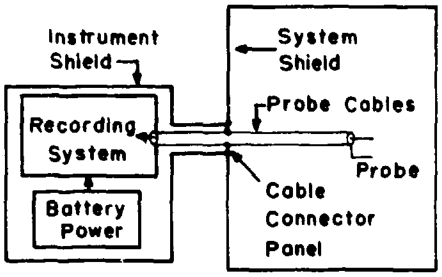
Point-of-entry testing provides information and definition of the coupled signals to critical portions of the system. Component damage and impact testing is commonly accomplished or based upon the energy content in a square pulse. Because many components are sensitive to waveform power spectra and polarity reversals, it is preferable to use waveforms which more accurately approximate the actual signals measured during the point-of-entry testing.

Test Point Accessibility

The critical points in a system that must be monitored in a test program are often internal to the system in a package that has limited space and access. Thus, to make measurements at these locations, some modifications must be made to the system. The use of circuit extenders is one technique employed for voltage measurements at hard-to-get-at critical points. However, the addition of extenders requires space that is not always available.

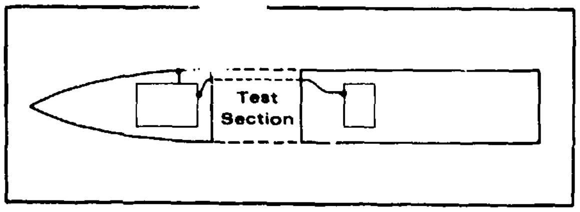
Limited space in a system often requires that the recording instrumentation be remote from the measurement point. With or without the use of adaptors, it is often necessary to cable the measured data to a location that is accessible during testing. Such a location could be a connector panel that is located in an individual equipment case or cabinet or can be on the external surface of the system. The routing of the cables and the shielding integrity of the connector panel should be accomplished in such a way as to minimize their effects on system performance during testing.

The recording system or nonconducting data transmission system can be attached to a cable connector panel by using a shielded enclosure as illustrated. The instrumentation box and system shields are constructed as one continuous shield with the system and instrument areas separated by the connector panel. The instrument box can also be used to contain a dielectric rod or microwave data transmission system.

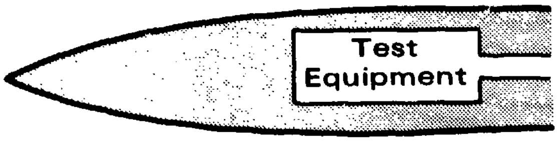


SHIELDED INSTRUMENT ENCLOSURE

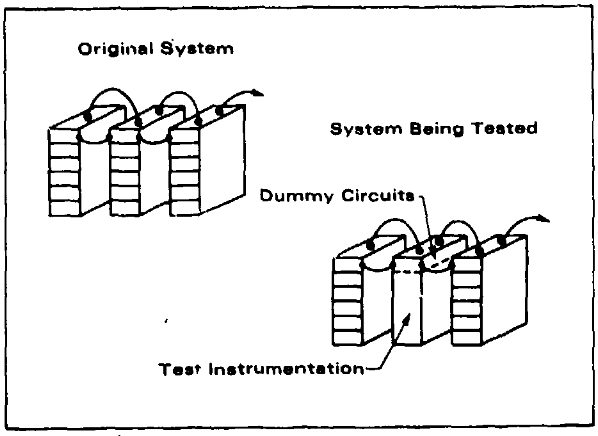
An alternate method to adding an instrumentation enclosure is to add space to the internal part of the system. Shown, in concept, is the addition of a section to a missile to contain test instrumentation. The section is added where the extensions to system cabling are a minimum (one cable shown). The test section can contain an access door. By keeping the system extender small compared to the total system size, the effects upon the EMP hardness or coupling to the system can be minimized.



Sometimes space within the system can be found for the measurement system by removal of some noncritical portion of the system. The most obvious place to find space is in some non-electromagnetic portion of the system such as a rocket motor bottle, for example.



At times it may be necessary to replace a portion of the system with dummy circuits so that instrumentation space can be found. The dummy circuits should terminate all wires that enter the circuit in their operating impedances. Particular care should be given to the circuit cases and ground circuits to ensure that the effects of the dummy circuits on the system performance are minimum. By replacing different portions of the system with dummy circuits, all parts of the system can be tested.



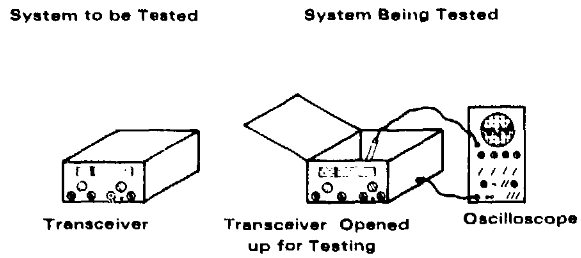
The alterations to the system to provide for test instrumentation also apply to modifications required for functional testing. For functional testing some operational mode must be simulated. If this mode is the in-flight performance of a missile, for example, some test hardware must be provided to simulate in-flight response of the missile. This type of simulation can either be mechanical or electrical and can require its own instrumentation to verify proper performance during test. This added "test set" must not be susceptible to EMP or alter the susceptibility of the system under test.

If the system under test is a portion of a larger system, then functional performance can only be checked by providing the missing portions of the system or their functional equivalents. A functional programmer is often required to provide the proper stimuli to test the system. This programmer can be remote to the system under test or it can be added to the system that is being tested.

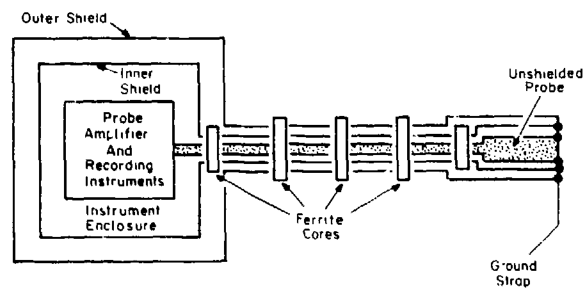
Precautions

The use of electronic equipment to make an EMP hardness or susceptibility measurement may alter the system susceptibility to EMP. If the measurement equipment is comparable in size to the system being tested, then the response in all likelihood is grossly affected by the measurement system. If the measurement equipment can be made small enough not to alter the electromagnetic fields and does not influence the system measurement point, then reliable measurements can be attained. Nonconducting transmission lines are often used to minimize equipment located in the vicinity of the system under test. The additional complexity and costs of using nonconducting transmission paths is not always warranted. Often, it is possible to use larger, simpler equipment if sufficient care is applied to how the equipment is positioned.

A measurement system can alter the measurement in several ways. First, the measurement equipment can be susceptible to the simulated EMP environment. This can be determined experimentally by observing the equipment in the environment. If the system is a probe measurement, then the lack of a response with no signal applied at the probe is verification that the measurement equipment is not susceptible to electromagnetic fields. For sensor systems, this test is usually accomplished with the sensor removed and replaced with its electrical equivalent. If the measurement equipment responds to the electromagnetic field, then shielding and other EMP hardening design practices must be applied to the equipment.



The susceptibility of the measurement system to EM energy is an indication of the capture of energy that exists in all conducting materials that are immersed in an EM field. Shielding restricts this captured energy to the outer surface of the equipment. The captured energy on the shield can be induced into the system being tested. However, this energy can be limited to the ground or reference side of the measured signal by using doubly shielded cable and instrumentation as illustrated.



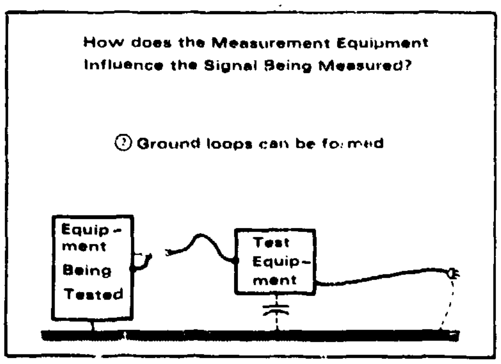
DOUBLY SHIELDED PROBE AND INSTRUMENT ENCLOSURE

The measurement system can also form unwanted ground loops that are coupled directly to the measurement point. These loops always exist but can be minimized by using very short leads from the probe to the point being measured. Remember, a voltage measurement requires two points to make the measurement and a differential measurement requires three points. The probe reference should be directly tied to the signal reference for single-ended measurements. If the signal reference is sufficiently stable so that the addition of other equipment references do not alter the signal reference, then the measurement recording equipment grounds and frames should be well bonded to this signal reference. Often the addition of the reference side (frame) of an oscilloscope, for example, will introduce unwanted currents into the signal reference. In this case, it is desirable to float the measurement equipment. Floating minimizes the capacitive coupling as well as ground loops from the measurement equipment reference to the system being tested.

Grounding of multiple instrumentation boxes should be at a common (single) point in the instrumentation system. Often, this is accomplished by the shields on the instrumentation cables. If these instrumentation cables do not all enter or exit the box at a single entry point, additional currents can flow over the box resulting in increased pickup internally. In some instances, multipoint grounding is unavoidable. In this case, the ground loops formed should be of minimal area. This can be achieved by grounding the instrumentation boxes with very short bonding straps and running the instrumentation cables in very close proximity to the grounding structure. To check for ground loop pickup, a grounding scheme or system which can be slightly altered with no discernible change in the data implies that the grounding system is not picking up unwanted signals.

Bonds and ground leads should be as low impedance as is feasible (i.e., large diameter, short length) to minimize the voltage drop in the reference side of the measurement.

Ground loops can also be caused through the test equipment power leads. The use of self-contained power sources is often warranted.



Probably the first EMP test performed upon a system will be to determine the effects of the test instrumentation and system test modifications upon the EMP susceptibility of the system. The techniques that are used for such tests are not easy to describe since many factors can alter the EMP susceptibility of a system. However, two general conceptual approaches are helpful to determine if the test results from the system and test equipment together are giving meaningful results.

Incremental addition is adding the instrumentation and modifications one at a time in as small an increment as possible to determine if any noticeable effect upon the system can be observed. This procedure then defines the effects of each addition on the system response.

An alternate approach is substitution. Substitution involves making test observations with and without the other instrumentation or modification of the system. Example, observe one test point while connecting or disconnecting other probes or grounding points.

Either of these two concepts requires that some initial instrumentation be added to the system to make the first observation. This instrumentation can be less complicated than the final instrumentation and, consequently, smaller in size. For example, instrumentation to make broadband time waveform recordings is much more complicated and larger than single-frequency-peak-reading instrumentation. Thus, if warranted, the effects of final instrumentation on the system can be investigated using CW, even though the instrumentation is designed for use with EMP-type measurements.

For data collection by any probe or sensor that is balanced (i.e., current probes, differential voltage probes, loop sensors, and balanced dipole sensors) unwanted pickup can be determined by making two data recordings, one of which is with the probe or sensor reversed. The reversal should result in a repeated signal magnitude with a polarity reversal for time waveforms or a 180° phase reversal for CW. Interference or pickup will show as the difference between the two data recordings. Often the magnitude of the interference is determined in data reduction so that some data should always be recorded with probes reversed, even though no appreciable interference is noticeable.

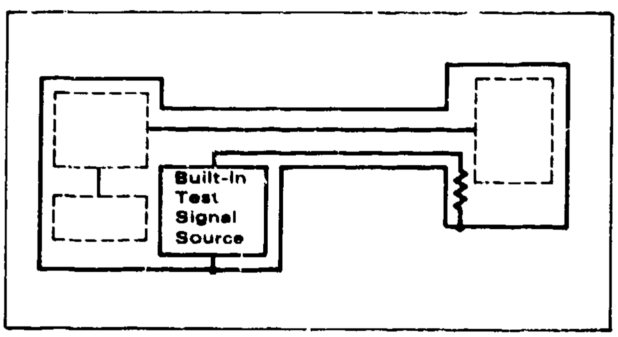
Quality Assurance Testing

Quality assurance testing encompasses a wide variety of testing techniques. At the component level, components can be checked to confirm that they perform as required prior to their inclusion in a final system. The actual testing technique used will depend upon the component and system. This form of quality assurance testing requires that the entire system has been tested so that the system specifications can be related to measurements on individual components. These measured characteristics at the component level can then be used to develop specifications for quality control. This same approach can be applied to equipments and subsystems.

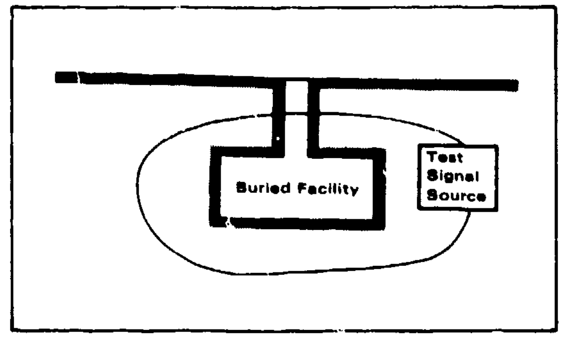
At the system level, quality assurance tests consist of tests to monitor the EM response and hardness of the system, and tests that can be performed to ensure system hardness over the system's intended life cycle. Quality assurance monitoring of a system will generally require some form of energy source to introduce a monitor signal into the system. Ideally, this source would be the simulated threat. Because of the complexity inherent in providing this threat, however, another source that can be related to the system hardness is often used.

One possible quality control test source could be a CW source that radiates at a few discrete frequencies. The resulting coupling to critical points could be determined using sensitive narrowband receivers while the system is active. For some installations, the CW source could be existing radio signals.

Quality control can be built into a system to test critical portions of the system. For example, cable drivers can be included or control wires added to conduits that are used to inject signals into critical cables.



The shielding effectiveness of a buried facility could be monitored by forming a conductive loop around the structure to provide an energy source.



The relationship between the system threat response using an EMP simulator and the quality assurance response using a simplified, non-threat waveform must be established during the system hardness testing. Knowing this relationship defines the quality control criteria for the monitor testing.

Some additional tests can be conducted during an initial test program that can replace future monitoring. These tests are generally confined to items that are suspected of showing deterioration with time. For example, if an access hatch can make a questionable RF seal, then a worst-case condition can be established for testing by leaving the hatch loose or open. If hardness is still maintained, then quality assurance to check the RF seal is not required. An open hatch may be unrealistic. However, the worst condition that the RF seal can possibly attain would be simulated for worst-case testing.

This concept during the system testing phase can eliminate many potential quality control measurements.

It should be emphasized that most systems are not static and will undergo modifications. The effect of these modifications can change the EMP hardness of the system and can be detected with a good quality assurance measurement program.

7.9 TEST PLANNING

Previous sections of this chapter have discussed the simulation techniques, the characteristics of EMP simulators and instrumentation necessary and available for performing an empirical susceptibility assessment of a system. This section will discuss the need for an organized system test plan, the overall test planning and scheduling, and the constituent parts of the detailed test plan. The discussion will be directed toward a system level test in a large scale simulator since this level of test planning is generally the most complex.

Pre-Test Planning

Pre-test planning is an absolute necessity in any test program. The time spent in the pre-test planning will be recovered many times over during the actual conduct of the test program.

It must be recognized that testing at the system level is but one phase of the overall hardness assurance program. The other phases of analysis, design, and laboratory testing all impact the test program. Therefore, test planning to coordinate phases should be initiated at the onset of the program.

A well thought-out test program will result in a minimum expenditure of time and cost for the system level test. Further, well defined test objectives will insure that the proper tests are conducted. The results of the system analysis and a bench testing program are additional important required inputs to the test plan since these inputs can, if properly executed, reduce the time required at the simulator facility.

System level testing requires bringing together several personnel teams. These teams include:

- Simulator operations team
- Instrumentation operations team
- Data processing team
- System operations team

To efficiently coordinate these teams in terms of preparing for and conducting the test with a minimum of lost time requires a well thought-out plan. The plan must spell out the who, what, when and where for each step of the test program.

A typical complex weapons system test takes about one (1) year to plan and organize. It is an iterative process which is usually conducted in three (3) stages. These stages are defined as: (1) General Program Plan, (2) General Test Plan, and (3) Detailed Test Plan. The stages (plans) have increasing levels of detail. Each stage is subjected to review and discussion. The objective of the review process is to marry the system with the appropriate facilities (simulator and instrumentation) to meet the objectives of the test.

General Program Plan

The first step in preparing for a system level test is the formation of a general program plan.

The general program plan must contain; as a minimum:

- General statement of the test objective
- Qualitative description of the system
- General description of threat criteria
- General description of the proposed tests
- Proposed test schedule
- Priority of proposed tests
- Organizations to be involved in the tests.

The general program plan must present a concise statement of the purpose of the test program. Often, it will include a list of primary objectives and a list of secondary objectives that are to be a achieved. Probably the most abused statement in the objectives of past test programs was: "To prove that the system could survive the EMP from a nuclear detonation." Since above ground nuclear testing is banned, it is impossible to prove that systems can survive their real use environment. Tests provide information which can be used in an analysis which estimates the survivability of the system in anticipated real environment.

The statement of the test objectives must be more specific. Example test objectives might be:

- To validate the EM coupling analysis for a structure
- To validate analytic predictions and estimates for coupling to deliberate antennas or to interior cables
- To aid the understanding of possible upset modes on major subsystems
- To verify that EMP hardening designs are sufficient for pro-

tection of mission critical systems, etc.

Having a well defined set of test objectives, a test program can be outlined for achieving these objectives.

A qualitative physical and functional description of the system to be tested must also be provided in the general program plan. The physical description must include the size of the structure(s) comprising the system, interconnection (configuration) of the structures, typical deployment, and identification of deliberate coupling sources. The functional description of the system must identify the operational modes and configurations of the system and the mission critical subsystems. In addition, the functional description should identify those functions that must be monitored to evaluate system performance before, during and after the tests.

The threat criteria should be identified in general terms. These threat criteria should be redefined in terms of the desired test environment criteria. The test environment criteria should define the test waveform, angles of arrival, polarization(s), initial test level and incremental steps desired, maximum field strength desired, and the relatability of the test environment to the threat criteria. The requirements for field illumination, direct drive, or combination thereof should be stated.

A general description of the proposed tests and the priority assigned to the various tests should be defined. At this stage, the test description need only identify the type of test, i.e., skin current or current density on the structure, interior cable sheath, bulk core, and/or individual wire currents, interior E and H fields, etc. Test points need not be identified at this stage, but the total number of data points for each test should be estimated. In addition, the number of system orientations, configurations, and operational modes to be tested should be identified. This will enable the total number of data points to be estimated.

The proposed test schedule should be identified. Estimates, based on test priority, of the time to be allocated to each test phase should be stated. The total time the system will be available for test and when it will be available should be specified.

The final aspect of the general program plan is identification of those contractors and government agencies who will take part in the tests. The responsibility of each group should be established. In addition, the support to be provided by each group, in terms of manpower, logistic, and instrumentation support should be discussed. Lead personnel who will constitute the test working group representatives from the various contractors and agencies should be identified. Administrative personnel who will work out administrative details, funding of the support groups, and logistics supply should also be identified.

Based on the data supplied in this plan, a test working group can be established to work out the ability of the participating organizations to support the test. Modifications to the test plan, presentation of alternative test plans, presentation of contingency test plans and establishment of final test schedules can be prepared. This general program plan should be submitted approximately six to ten months prior to the anticipated test date.

General Test Plan

The general test plan is the second iteration in the overall test planning process. It is a more detailed statement of the tests to be performed, and a complete technical description of the system to be tested. It should include as a minimum, the following:

- Precise statement of the objectives of each test to be performed
- Description of the test approach to be utilized to meet the test objectives
- Environment levels and orientations for each test, i.e., simulator requirements
- Complete technical and functional description of the system (test model)
- Instrumentation requirements
- Data reduction and processing requirements

A precise statement of the objectives of each test to be performed must be prepared. It should identify the reason for the test, the data to be obtained, and the use that will be made of these data.

The test approach to obtain the data identified in the test objective must be specified. The test approach selected will depend on the size and complexity of the system to be tested, constraints such as whether destructive testing is permissible, availability of environment simulators to produce the environment required, and availability of instrumentation to measure the expected signal levels. The test approach may utilize any of the test concepts discussed earlier in Section VII. The test working group will direct the test approach based on the general program plan and information on available simulators.

A description of the environment characteristics for each test objective to be accomplished must be defined. This description must include field polarization, field level(s), angle(s) of arrival, volumetric coverage required, general waveform characteristics (rise time and fall time or ε folding time). Requirements for field mapping in the vicinity of the test system, and monitoring of shots should be defined.

A complete functional description of the system must be presented. This should include operational modes, identification of mission critical elements, and functional checks to evaluate system performance. Special safety precautions should also be specified if damage or upset of an operating system can result in potential hazards to the test personnel. If these safety measures require special equipment such as shrouds for laser systems, safety goggles, safety blocks, etc., these should be identified.

The technical description must be sufficiently detailed to provide the required information for placing the current and voltage probes. This may involve equipment location plan and perspective drawings, circuit block diagrams and schematics, identification of desired test points, and means of access to test points. If special break-out boxes or equipment modifications are required, these should be specified. If these additions or modifications can effect the system response, they should be included, if possible, in the pre-test analysis of the system. Constructional features, especially for EM shielded portions of the system, should be provided. Many times, photos can be used to great advantage to present this information.

The results of the pre-test analysis and any in-house laboratory testing should be utilized to the utmost. These pre-test data will define the anticipated signal levels and frequencies, and predict the potential for subsystem/system damage or upset. Pre-test analysis can also aid in defining test points, potential points-of-entry into the system, expected shielding characteristics of various portions of the system, etc.

The results of the pre-test analysis and in-house laboratory testing are an indispensable aid in determining the test instrumentation requirements. The test instrumentation requirements should be specified in as much detail as possible. If standard commercial probes, sensors, data links, and signal conditioning equipment are to be utilized, specification by commercial part number may be adequate. However, since there is usually more than one type of commercial, or government instrumentation that are equivalent, it is often a good idea to include the data sheet information to describe the instrumentation. The general information necessary on probes is bandwidth, sensitivity, input and output impedance, dynamic range, common mode rejection ratio if applicable, shielding requirements if applicable, and physical characteristics (primarily size, required diameter for current probes, etc.)

Required bandwidth of data telemetry and data recording systems must also be specified. Gain requirements, dynamic range, and levels to be recorded can be determined from expected signal levels and sensitivity of recording equipment. Data recording requirements also include time base or sweep rates for time or frequency waveforms. Requirements for multiple recordings per test channel should also be specified. The need for custom designed or modified instrumentation requirements should be indicated including specifications, circuit diagrams and fabrication drawings. This applies also to break-out boxes for attaching current and voltage probes.

Data reduction and processing requirements should be described. If quick-look data is desired, it should be indicated in the test plan. Other data processing requirements might include digitization of data, time-to-frequency transforms, computer analysis and printout of peak amplitudes and dominant resonant ring frequencies, scaling to threat criteria if low-level simulators are used, etc.

The general test plan should be distributed to all participating organizations approximately four (4) to six (6) months prior to the test date.

Detailed Test Plan

The detailed test plan represents documentation for the performance of the test. The results of the review of the general program plan and the general test plan by a test working group and an administrative group must be reflected in the final document. As a minimum, it must contain the following information:

- Specific objectives of the overall test program and each phase of the test program

- Detailed description of the test model noting all differences (including modifications incorporated for test instrumentation purposes) from the standard model

- Complete description of the test environments, and simulator characteristics for each phase of the test program

- Supporting drawings and diagrams of the system, instrumentation, component wiring diagrams, functional diagrams, etc.

- Functional procedures for system performance checkout and evaluation during/after each phase of the test program

- Complete description of the system configuration, orientation, and operational modes for each phase of the test program

- Detailed description of all instrumentation including purpose and characteristics of each probe and sensor.

- Detailed test and calibration procedures

- Complete description of all data to be obtained for each phase of the test program

- Complete description of the data flow and reduction procedures to be used during and after the tests

- Detailed schedule that specifies day-by-day measurements and documentation delivery requirements

- List and format of all logs and summary documents to be generated on the test program and the organization responsible for preparation and submittal of the document.

Administrative information must also be submitted concurrently with the technical description. This information may be supplied as a part of the detailed test plan or as a separate document. This information identifies the responsibility and authority of each organization participating in the test program. As a minimum, it should include the following information:

- Each organization's relationship and authority

- Each organization's responsibility for providing support in the area of:

  a. equipment and instrumentation

  b. administrative and test personnel

- Security and safety requirements

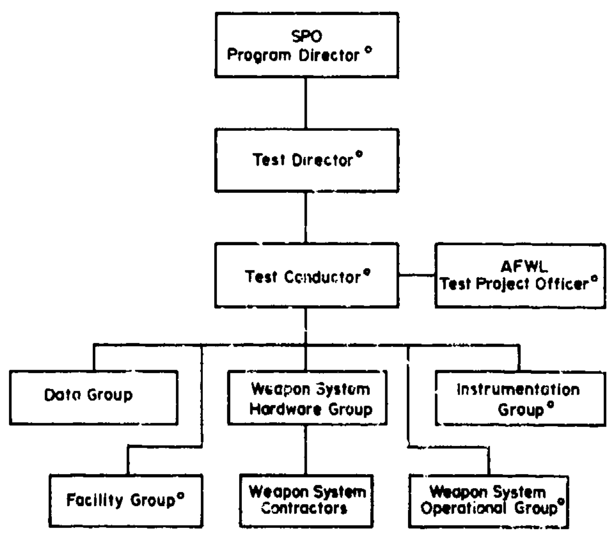
- Logistics

The detailed test plan elaborates on the general test plan. The revisions to the general test plan required as a result of the reviews and meetings associated with the test planning function, and any new material in these areas, must be incorporated into the detailed test plan.

In addition to the material included in the general test plan, are detailed tests and calibration procedures. These procedures should be step-by-step technician oriented instructions for instrumentation setup. Means of identifying data quality (such as noise determination shots and probe reversal) should be documented. The instructions should specify how often noise checks and calibration must be repeated.

These step-by-step procedures generally do not have to be documented for the simulator operations team. Simulator operations procedures are well established in most cases and specially trained personnel familiar with these procedures are the only persons authorized to operate the large scale simulators.

A typical organization chart for a test program to be conducted at the Air Force Weapons Laboratory (AFWL) is illustrated.



° Indicates Mandatory Test Working Group Membership

TYPICAL TEST ORGANIZATION

The Test Director is usually from the prime contractor's organization if the objective of the tests is to provide design information. If the tests are certification type tests, he may be a representative from the System Project Office (SPO). The Test Conductor is usually from the organization responsible for the simulator and instrumentation facilities. The remaining groups can be drawn from any of the organizations participating in the test program. The mix of these groups will depend greatly on the nature of the test program and system under test. The authority and responsibilities of each individual or group must be identified to result in an efficient, orderly test program.

The number of test support personnel from each of the participating organizations must be identified. Detailed job functions for each group, and in some cases, individuals, must be prepared. In this way, coverage for all initial test functions is assured.

Security and safety requirements must also be identified as well as the cognizant person in charge of these activities. Generally, one person from the prime contractor and one from the testing organization is given responsibility for these duties. Security requirements include classification of hardware, classification of data, and physical security to protect the system from tampering. Safety requirements include potential shock hazards, personnel safety during checkout if potentially hazardous systems are involved, protection of fuel subsystems if vehicles are fueled, removal or safing of ordnance, etc. If required, protective devices and their suppliers must be specified.

A cognizant individual or group should also be identified for all logistic functions, especially if non-durable provisions are necessary. This includes fuel for operating the system, film for instrumentation cameras, documentary photography, special targets for weapons systems, etc.

The detailed test plan must be submitted between one (1) and two (2) months prior to commencement of the tests.

Detailed Test Plan Outline

An outline showing some of the important tasks to be performed in the overall test program is provided here for guidance and to indicate the salient features of a test program. Every test program will not necessarily contain all elements listed.

Detailed Test Plan

(Typical Outline)

1.0 Introduction

- 1.1 Purpose
- 1.2 Test Objectives
- 1.3 General Description of Test Program

2.0 Test Program Management and Responsibilities

3.0 System Description

- 3.1 Physical
- 3.2 Functional
- 3.3 Identification of Mission Critical Element

4.0 Pretest Analysis

- 4.1 Coupling Prediction
- 4.2 Prediction of Signal Levels
  - 4.2.1 Cable Sheath Currents
  - 4.2.2 Terminal Voltages and Currents
- 4.3 Prediction of Damage/Upset Susceptibility
- 4.4 Identification of Points of Entry
- 4.5 Identification of Critical Circuits
- 4.6 Identification of Worst-Case Coupling

5.0 General Test Requirements

- 5.1 Simulator Characteristics
- 5.2 Instrumentation Requirements
- 5.3 System Modification Requirements
- 5.4 Special Test Equipment Requirements
- 5.5 Operational Safety Requirements
  - 5.5.1 System
  - 5.5.2 Simulator
- 5.6 Support Requirements
  - 5.6.1 Maintenance Spares
  - 5.6.2 Special Fuels for System
  - 5.6.3 Supplies
  - 5.6.4 Test Equipment
  - 5.6.5 Data Processing
  - 5.6.6 Security

6.0 Test Procedures

- 6.1 Scope

6.2 Operational Checklists
6.3 Performance Verification
6.4 Detailed Measurement Procedures

6.4.1 External Fields
6.4.2 Exterior Surface Currents
6.4.3 Internal Fields
6.4.4 Cable Sheath Current
6.4.5 Cable Core Current
6.4.6 Terminal Voltage/Current
6.4.7 Terminal Input Impedance
6.4.8 Upset Measurements

7.0 Test Schedule

8.0 Documentation

8.1 Test Log
8.2 Test Data
8.3 Test Reporting

REFERENCES

"DNA EMP Preferred Test Procedures." DNA 3286H, Defense Nuclear Agency, Washington, D.C. 20305 (under Contract DNA001-75-C-0074 to IIT Research Institute, Chicago, Illinois 60616), Revised February 1977.

"Engineering Techniques for Electromagnetic Pulse Hardness Testing," DNA3332F, Defense Nuclear Agency, Washington, D.C. 20305 (under Contract DASA01-71-C-0087 to Stanford Research Institute), September 1974.

Electromagnetic Wave and Radiating Systems, E.E. Jordan, Prentice-Hall, Inc., 1950.

"Electromagnetic Pulse Instrumentation Handbook," Air Force Weapons Laboratory, Kirtland AFB, N.M., AFWL EMP Measurement 2-1, October 1971.

"Electromagnetic Pulse Sensor Handbook," Air Force Weapons Laboratory, Kirtland AFB, N.M., AFWL EMP Measurement 1-1, June, 1971.

SECTION VIII

APPROACHES TO VULNERABILITY ASSESSMENT AND SYSTEM HARDENING

8.1 INTRODUCTION

Hardening of a system against the EMP resulting from a nuclear weapon detonation is a complex problem. Previous chapters have discussed the environment generation, the potential effects on systems, system protection, and the analytical and experimental tools available to perform a hardness assessment of the system and verify the hardness design. The available knowledge, and the present capability in each of these areas is limited and a degree of uncertainty still, and probably always will, exist.

EMP hardening is a system problem, that is, the approach and hardening implementation are highly system dependent. Further, hardness assurance is a total system life problem which must be considered during the hardening design phases. The design must be cost effective; this results in additional considerations (tradeoffs) of cost, performance, operational considerations, logistics and maintenance, and field modifications.

The state-of-the-technology in EMP hardening is not sufficiently advanced to provide detailed hardness standards and specifications. There is no single standard plan for EMP hardening. Each system manager must consider the EMP hardening in the light of his system peculiarities, needs, feasibility, time and cost, facilities and contractors.

8.2 GENERAL APPROACH TO EMP HARDENING

EMP hardening must be considered from the onset of the program through the operation/maintenance phase of the program as illustrated. At program onset, the threat scenario and the system mission must be considered. These inputs provide preliminary information for determining what portions of the system must be hardened and the total hardening requirements. Usually, because no detail is available on the system, this implies performing a "worst case" analysis. Each of the threats identified in the threat scenario defines an environment the system must withstand which, along with the defined operational modes of the system, identifies a set of initial system design criteria.

Design criteria for protection from EM effects other than EMP should be interjected at this time. Cost effective hardening for any EM environment requires maximizing performance at the minimum cost. If costs can be shared, the costs for the entire EM hardening effort are minimized. These cost savings occur both in the design and production phases of the program. Further, it is desirable to minimize the system complexity by insuring redundant hardening for the various EM disciplines is eliminated. Redundant hardening generally results in unnecessary components which will have an adverse effect on system reliability and maintainability.

The resulting EM control plan will identify the concepts by which hardening is to be achieved. It will provide, for example, the grounding philosophy (single or multipoint, etc.), the cabling philosophy (wired vs. chassis signal and power returns, cable shielding, etc.), the shielding approach (overall envelope versus shielded equipment bays, etc.), to name a few. These concepts are translated into a set of design criteria to be applied at the system level and the subsystem/equipment levels, referred to as "allocation of the hardening requirements." A second output is a hardening design plan which stipulates certain aspects of the system and subsystem designs which must be controlled very tightly to achieve compatibility. Thirdly, it defines how the system/subsystem designs are to be verified to minimize the need for redesign. This verification will be performed on prototype portions of the system since the complete system is usually not available. They are generally laboratory type tests at the component level up to the subsystem level and are usually direct injection or cable driver tests. The results of these subsystem/equipment/component tests or analyses are inputs to the design evaluation task. System considerations are also inputs to the design evaluation. The various elements of the design are conditionally verified at this time.

During the prototype development phase, verified designs are implemented at the system level. At this time, system tests in a suitable simulator are planned. These tests should be designed to define the worst case system configuration (environment and coupling). Diagnostic tests should be conducted to verify the predictive analysis with regard to anticipated cable currents and terminal voltages and currents. If a threat criteria level simulator is available, damage or upset verification tests can also be performed. Numerical analysis plays a role during this phase to relate the test waveform to the threat, scale system responses to the specification criteria, relate test data from EMI and TEMPEST tests, assess system performance, etc.

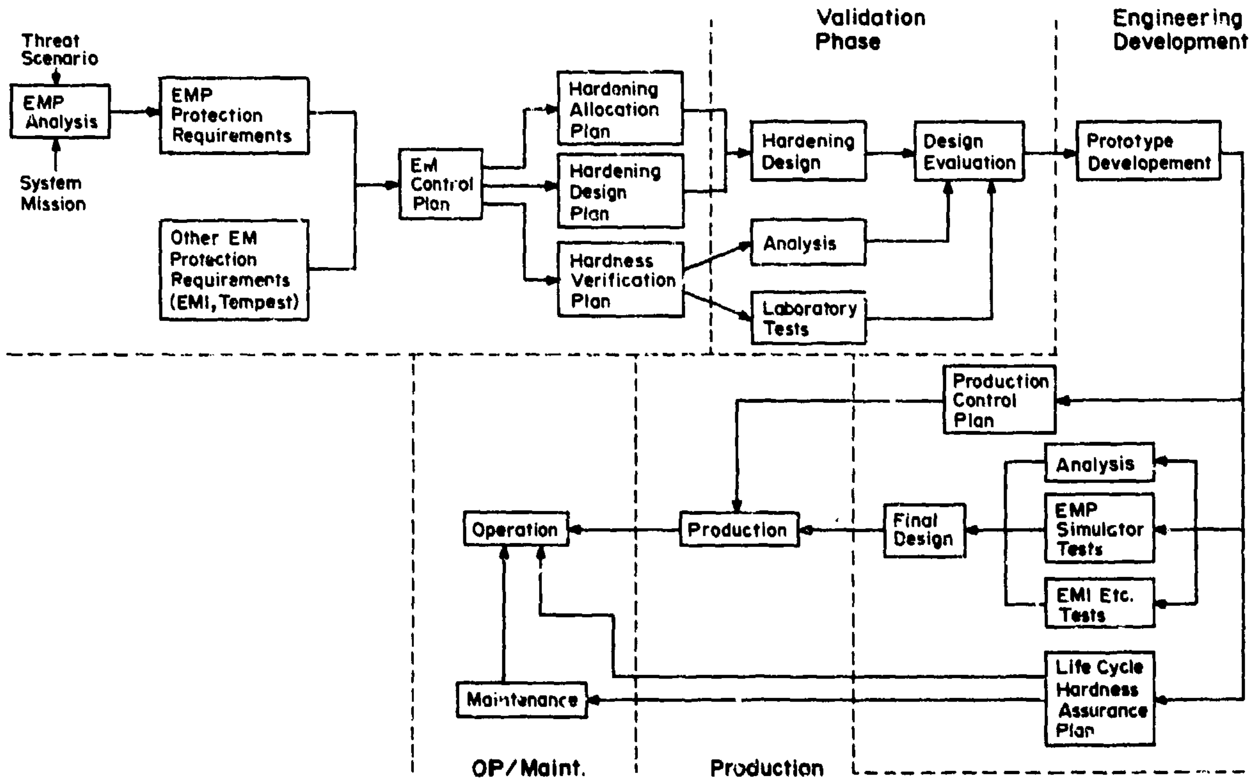
During the prototype development phase, the production control plan and the hardness continuation plan should be prepared. The production control plan must provide for controls at all levels, component through final system. Procurement specifications must be prepared which include parts control specifications and incoming inspection and control. Detailed production drawings, test fixtures, specifications, and production line monitoring procedures must be prepared for in-house fabricated portions of the system. Final assembly and monitoring of the elements of the system and system level certification tests must also be specified.

The hardness continuation plan must provide the necessary inputs for the user (system operator, depot, field maintenance, etc.) to maintain the EM hardness designed and built into the system. It must specify maintenance procedures and schedules for the EM hardening elements.

Where parts control is essential to the hardening level, a list of approved parts containing sufficient detail to assure correct replacement must be prepared. Procedures and schedules must be identified for all levels of maintenance. Operational procedures critical to the EM hardness level should be thoroughly described.

The final aspect of hardness continuation is the control of retrofit or future modifications of the system. An EM control program is as essential in this phase of the system life cycle as the original system design. If system modification controls are not specified in the same degree of detail, the overall system hardness can be compromised.

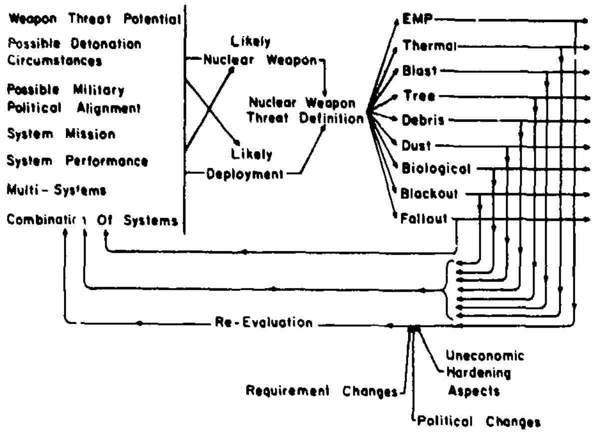
The discussion above is only a general outline of hardness considerations throughout program development. Real programs must be tailored to meet the specific needs, facilities, scheduling, and resources of the specific system under development.



SYSTEM LIFE CYCLE

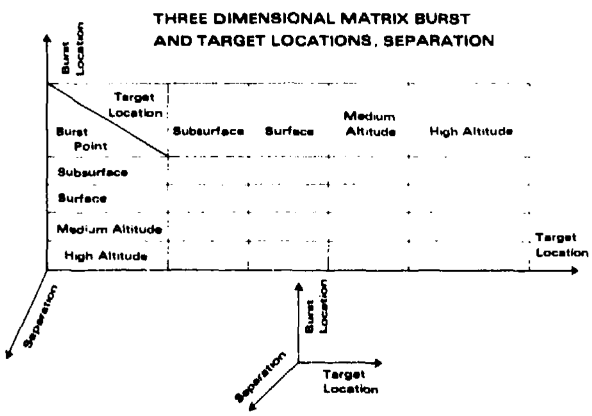
8.3 THREAT/ENVIRONMENT CRITERIA DEFINITION

For either planned or existing systems, the EMP environments must be developed in context of the performance and mission requirements of the system. The weapon threat, detonation circumstances, system configuration, system operational requirements, etc., must be identified.



Threat Scenario

The threat scenario must consider all aspects such as the type of weapon, yield, and likely targeting circumstances. The weapon type, yield, and burst location will define the EMP fields and waveform and levels of all other nuclear weapons effects. Inserting the identification of the possible target locations, the nuclear environment seen by the system can be determined. All possible combinations of importance can be described in the form of a burst/system location/separation matrix.



From this matrix and available data on the weapons effects (blast, thermal, shock, radiation, EMP, etc.), as a function of range from the burst point, all environments can be specified.

System Mission

In addition to specifying the threat scenario, the system mission and operational requirements must be specified. The operational requirements include whether the system must function through the event or only post event. This strongly impacts whether upset is a serious problem or not.

Further, it is necessary to specify the type of critical system elements (i.e., manned or unmanned, employment of digital computers or equipments, operating levels, etc.), and the system configuration (size, coupling structures, etc.). This system definition has a major impact on the required hardening design.

Nuclear Hardening Requirements

With a knowledge of the threat scenario and the system mission and operational requirements, the hardening specifications can be developed. The concept of "balanced hardness" must be employed.

To implement "balanced hardening" all nuclear weapons threats and survivability criteria must be considered. For example, if unprotected personnel are required for operation of the system, the survivability of the crew to radiation is usually the limiting factor on system survivability. Any one of the weapons effects can be the limiting criteria, and it makes no sense to harden a system to levels of EMP, etc., if it cannot be expected to survive some other nuclear weapon effect.

Information is available in the form of examples of other systems design and survivability criteria to help in this type of tradeoff study. These environmental tradeoff studies must be performed for each potential circumstance identified in the previous matrix. Specifications derived in this manner usually identify the worst case EMP that the system must survive for a near-surface burst and, in addition, usually include the high altitude burst case as a second set of EMP criteria. The high altitude burst case is usually included since, for most systems (i.e., surface and near surface systems), there are no other nuclear weapons effects associated with this burst location.

8.4 SUSCEPTIBILITY/VULNERABILITY ASSESSMENT

Assessment of the susceptibility and/or vulnerability of a system must begin with as detailed a technical description of the system as possible, identification and estimate of the EM coupling to the system, and an evaluation of the effects of the coupled energy on critical elements of the system. To perform this assessment requires adequate mathematical and experimental tools, and system models for analysis which are sufficiently detailed to be consistent with the various uncertainties of the problem.

System Description

The description of the system may vary from design concepts to detailed technical characteristics depending on the phase of development of the system. If the system is in the conceptual design phase, far less detailed information will be available than if the system is already operational.

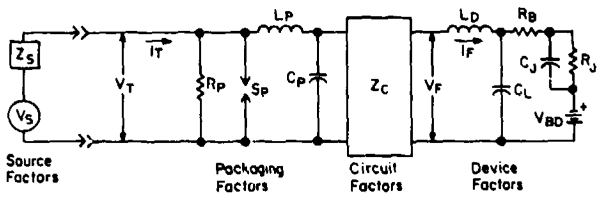
A physical description of the system is essential to performing the coupling analysis. This must include the layout of the system (number of elements comprising the system, inter-connection of these elements, etc.), and the physical size and shape of these system elements. In addition, any deliberate antennas must be identified. Characteristics of importance are size of the antenna (length, diameter, etc.), operating frequency, type (aperture, parabolic dish and feed, whip, dipole, long wire, etc.), and location (height above ground, supporting structure, etc.). In the case of new systems design, this information may have to be based on experience and the required performance characteristics. If options are available, each of these options should be described. Whenever possible, an attempt should be made to identify ports of entry for EM energy into the system. These may include such things as known apertures, hatches, access ports, ventilation ducts, and cable penetrations.

For existing systems, the problem of describing the system is greatly simplified. In this case, detailed drawings and technical specifications are available. Further, an actual system installation is available for scrutiny to identify ports of entry in the form of apertures or penetrations. Often experimental data are available on the technical performance characteristics of the system. Data may also be available on the EMI response aspects of the system which can provide excellent guidance to its EMP susceptibility characteristics.

Detailed circuit characterization of the critical elements of the system, especially interface circuits, is also required to perform an in-depth S/V assessment. Even if detailed circuit drawings and technical specifications are available for the system, the response characteristics and exact values of the circuit components are not adequately described. Often these component characteristics can only be obtained through a measurement program. In the case of systems which are in the conceptual design phase, the component characteristics and values can be estimated based on data sheet information, by measurements on typical or generic type components, or from existing data bases. The type of component data required includes frequency response characteristics and damage constants for both passive and active components. Logic levels, noise margins, etc., must also be known for digital circuits to perform an upset assessment of the circuits.

Electrical Overstress Assessment

The elements of an electrical overstress assessment are illustrated in the following figure. The source factors are defined by the EM coupling to the system. These factors depend on the coupling source pick-up and represent the driving function for the terminal voltages and currents.



ELEMENTS OF AN ELECTRICAL OVERSTRESS ASSESSMENT

The packaging factors are also system dependent. These factors include the metallic enclosure structure and cable termination at the points of penetration into the system.

The circuit factors include all passive components which determine the circuit response (transfer function) and determine the energy, voltage and/or current applied to the active component in the circuit. These factors must include the stray effects which can significantly alter the circuit response at high frequencies.

The device factors are those that are associated with the active devices in the circuit. Of primary importance are those devices which interface directly with the coupling sources.

There are several approaches to performing an S/V assessment. These approaches vary in the level of detail of either the analytical, experimental or combined evaluation. At the extremes are: (1) a "worst case" analysis, and (2) a "rigorous" analysis.

"Worst Case" Analysis

A "worst case" analysis is generally employed to obtain a first cut (or ball park) estimate of the system susceptibility and vulnerability. It usually utilizes simplified geometric coupling structures, a simplified environment waveshape, and a simplified electrical model of the terminal device impedance.

This type of quick-look analysis is particularly useful to determine whether or not a potential EMP problem exists. If a problem is apparent, this approach provides a high safety margin hardening criteria. Also, worst case waveforms for laboratory tests can be identified prior to full scale tests. Early isolation of potential problems enables design changes in circuit layouts, system groupings, etc., and provides a framework for a more complete analytical approach as required.

The quick-look energy comparison approach involves seven major steps:

1. Establishment of performance requirements:

   The first step is to determine the mission and performance requirements of the system and the significance of EMP-induced malfunctions.

2. Identification of susceptible components:

   The second step is to identify components susceptible to EMP degradation. In many cases, the degradation threshold is specified in terms of the minimum joule energy required to cause a malfunction. This often must be determined empirically on representative components.

3. Determination of representative EMP collectors:

   This is best done by actual visual inspection of the system, if available, or examination of the blueprints and system layout. For new systems, conceptual designs must be used. Simplified geometric models of the coupling structures must be developed. Effective height and source impedance must be estimated.

4. Development of simplified waveforms:

   From the specified nuclear weapon threats simplified EMP waveforms can be developed.

5. Development of simplified circuit impedance:

   Impedance match between coupling sources and circuit under study determines energy transfer.

6. Calculation of the worst-case energy:

   Based on the worst-case orientations of the typical EMP collectors, the maximum possible energy collected is developed.

7. Comparison of the worst-case energy and minimum degradation:

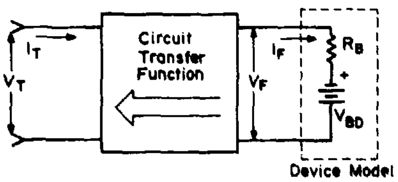
   As the final step in this worst-case type of quick-look analysis, the maximum possible collected energy is compared with the minimum energy required to degrade the various components. If the total energy available exceeds the minimum degradation energy, then a more detailed study, analysis and testing should be considered. A fair presentation is to depict the results on a logarithmic chart which indicates the wide range of energies available and associated uncertainties with the degradation thresholds.

Detailed Vulnerability Analysis

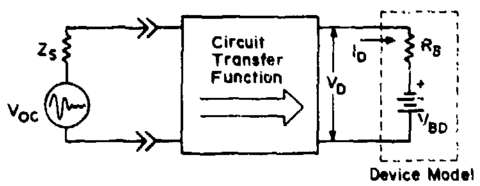
A detailed (rigorous) vulnerability assessment requires development of analytical models which define the response of the system. There are two generic approaches which can be applied: (1) the circuit inherent hardness estimate, and (2) the circuit vulnerability estimate. The circuit inherent hardness approach utilizes component failure data or circuit upset data and translates these device terminal voltages and/or currents to circuit terminal voltages and currents required to produce the response through the circuit transfer function. These "failure" terminal voltages and currents are then compared to the coupled voltages and currents to determine the margin of safety or protection requirements.

The circuit vulnerability estimate begins with the source (coupling estimate) in terms of an open circuit voltage and equivalent source impedance as the driving function. These data are translated via the circuit transfer function to determine the device terminal voltages and currents which are then compared to the failure or upset thresholds.

GENERIC APPROACH FOR DETERMINING CIRCUIT SPECIFICATION VULNERABILITY INHERENT HARDNESS



a) CIRCUIT INHERENT HARDNESS ESTIMATE



b) CIRCUIT VULNERABILITY ESTIMATE

V_{BD} - Device Breakdown Voltage
R_B - Device Bulk Resistance
Z_s - Source Impedance
V_{oc} - EMP Induced Open Circuit Voltage

The accuracy of either approach is dependent on how well the system can be modeled. The variables associated with the analysis are the coupled waveform, the complex impedance of the passive elements, the complex impedance of the active elements, the active element response (damage or upset model), the complex source impedance and, if experimental techniques were utilized to measure these data, the waveform used.

As stated previously, simple geometric models are usually employed for determining coupling to the structure. Simple models can be used to obtain exterior skin currents, cable sheath currents, and antenna currents. The time domain response waveforms are usually of the form of damped sinusoids or a combination of damped sinusoids depending on the coupling element resonances. Shielding estimates to obtain shield currents or unshielded wire currents on internal cables can also be performed using simple models. Terminal voltages and currents are determined utilizing the transfer impedance of the shielded cables. Computer codes (analyses) are available to determine the source impedance for coaxial or multiconductor cables. Lumped parameter circuit models have been developed for simple antennas which are directly usable in circuit analysis codes to provide the terminal voltages and currents. For existing systems, these terminal voltages and currents can also be obtained experimentally.

To perform an analytical assessment, models of the passive elements in the circuit must be developed which characterize element response for normal and extreme signal levels. High signal response models are required which will predict failure as a function of electrical overstress amplitude and duration. These models will predict the effectiveness of the passive components in protecting the active devices. The model topology must be simple enough to permit cost effective analysis but must include all parasitic R's, L's, and C's which significantly affect the circuit response. These model parameters usually must be measured experimentally.

Damage and response data have been documented from test programs for carbon composition, wire wound, metal film, and metal oxide resistors. The only analytical damage assessment model currently available is based on a multiple linear regression analysis based on the test data. Subsequent work indicates a physical assessment model is achievable.

Well documented test data is available for both electrostatic and electrolytic capacitors. Response models have been formulated for electrolytics but simplified parameterization procedures are incomplete. The only assessment model available is an empirical model which relates the transient breakdown voltage to the rated dc breakdown voltage.

Damage in inductors has not usually been a consideration. Response models have been empirically derived using network analyzers. It must be remembered that inductors behave as capacitors at high frequencies.

Damage and response models are also required for active devices in the circuit. Generic circuit response models for semiconductor components are available in most cases. These are equivalent circuit models which include the parasitic elements and current and voltage generators. They are large signal models, that is, the non-linear characteristics are included in the model.

Methods for obtaining damage models for use in vulnerability assessment for semiconductors include:

- Destructive testing
- Existing data banks (SUPER SAP II)
- Specification sheet data
- Non-destructive measurements
- Generic component category for which data exists.

These methods are documented in the handbooks.

Having developed correct models, a reasonably accurate assessment of system vulnerability can be obtained. The more accurate the model, the better the assessment. The most practical and cost effective approach is a combination of a analysis and test methods. Care must be exercised in the combined approach since empirical component failure data are most often obtained using simplified driving waveforms, a square wave being the most common. As a result, there is not a direct correspondence to the actual or predicted waveforms. Conversion factors must be developed based on a convolution integral solution. Conversion factors have been documented in the handbooks.

INDEX